IBA

TECHNICAL REVIEW

17

Developments in Radio-frequency Techniques

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17 Developments in Radio-frequency Techniques

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Engineering Research

by Baron Sewter
Assistant Director of Engineering
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Benjamin Jowett, when Master of Balliol, believed that research was a mere excuse for idleness. Nearer the mark is the view that the efficiency of a research laboratory is that fraction of its published output of which it could be said that it would have made a significant difference if the work had not been done. It was Churchill who said: 'It is always wise to look ahead, but difficult to look farther than you can see'.

IBA Technical Review Vol. 17 reflects some aspects of the applied research and development carried out in our Radio-frequency Laboratory: the work and thinking that led to the uniquely successful SABRE aerial on Alderney; the continuing work on the reduction of off-set co-channel interference by comb filtering at a fixed frequency; the detailed investigation of the troublesome effects of the noise output and jitter of frequency synthesisers; the painstaking development of precise yet flexible instrumentation for the field engineering of large transmitter networks; and a novel and improved demultiplexer for stereo and three-channel transmission systems.

All illustrate a vital aspect of modern r.f. technology: the sophisticated use of electronic circuit elements to minimise or overcome the problems imposed by the fundamental limitations of the r.f. medium—the thresholds of noise and jitter, the ways in which r.f. propagation falls short of the ideal because of super-refraction in the lower troposphere and the seemingly inevitable multipath and fading. One day perhaps guided wave propagation along optical fibres will bring relief from such limitations, but for the foreseeable future the vast majority of viewers and listeners will continue to depend on the free, if capricious, radiowaves at frequencies from MF to SHF.

The work also illustrates that, to the practising engineer, the distinction too often drawn between 'pure' and 'applied' or 'fundamental' and 'practical' research is largely specious. As that very practical engineer, Peter Baxandall, pointed out:
'It is, surely, a bad thing that there should tend so often to be a schism between practical designers on the one hand and 'university academic types' on the other, when a combination of attitudes—which fortunately sometimes occurs—can so greatly enrich both...

'It seems to be sometimes overlooked that engineering is logically concerned with *how to make things and solve practical problems*... the path along which the competent engineering designer travels in order to end up with a first-class design, whose final theory is properly understood and in which the effects of component tolerances are known, is often very different from that implied in many text books and published articles... more discussion of such practical matters as parasitic oscillation, earthing techniques, minimising hum etc.... would help less experienced design engineers to become aware of some of the things that have been found out the hard way by others... millions of pounds must be spent annually by industry in sorting out such "mundane" matters.'

Of course pioneering can be a painful and costly experience. Not everyone may be quick to see the potential of new ideas. When, for example, in 1943 British engineers described a new technique they had called 'frequency synthesis', the IEE discussion period was largely taken up by condemnation of this new addition to the terminology—and it was many years before the technique itself received the attention it deserves! Past events, past ideas cast long shadows both before and behind them.

Again, it is popularly supposed that, with almost a century of work since Popov raised his first receiving aerial, there can be little of this seam left to be worked. Yet in the past decade digital phase shifting, allowing several polar diagrams to be shaped simultaneously under the control of a digital computer, has opened a whole new and exciting adaptive technology that integrates electromagnetic theory, low-noise amplifiers, diode digital phase shifters, ferrite components and phase-lock-loops into the burgeoning world of the r.f. engineer. A world into which, by introducing the concept of optimisation by signal-to-noise ratio, the IBA team has made a real and lasting contribution.

This Review contains a selection of papers that illustrates, with unusual clarity, that practical broadcast engineering research and development, aimed squarely at benefiting the viewers and listeners, involves working at or beyond state-of-the-art technology; that there is still rhyme and reason for innovation engineering based solidly on a sound study of fundamental theory and practice.

PAT HAWKER has been with the IBA Engineering Information Service since 1968. After wartime special communications work for various Intelligence services, he was engaged in the publishing, editing and writing of technical books and periodicals, and remains a frequent contributor to the technical press. Books in print include *Amateur Radio Techniques* (7th ed.) and *A Guide to Amateur Radio* (18th ed.)

Radio-frequency Developments—An Introductory Survey

by Pat Hawker

Synopsis

A Radio-frequency Laboratory, within the broadcasting environment, is concerned with the application of new circuit and signal-processing techniques to overcome limitations on the picture and sound quality of television and radio transmissions. Modern technology has significantly increased the sensitivity of receivers but paradoxically has resulted in the effects of oscillator noise and jitter, particularly that of frequency-synthesised sources, becoming more important. Inherent distortion of vestigial sideband transmissions when demodulated by a simple envelope detector has led to the use of synchronous demodulation in receivers and test equipments. Much effort has been expended by r.f. engineers in seeking to minimise co-channel interference (CCI) brought about by trans-horizon anomalous propagation of UHF signals and this section describes some of the mechanisms that result in such propagation over sea and land paths.

The use of circuit and aerial elements to combat CCI and multipath propagation is considered. A brief account is given of the development of the MSC surround system and the requirement for an improved demultiplexer for two- or three-channel transmissions.

The role of a Radio-frequency Laboratory within the broadcasting environment is both evolutionary and revolutionary. Primarily, it involves the application of modern circuit and signal processing techniques to overcome problems, many of them long-standing, that impose limitations of technical quality and/or coverage on television and sound-radio transmission and reception. The aim is not only to improve and extend the performance or operation of existing broadcasting systems but also to investigate and develop new systems, both for direct-broadcast satellites and for terrestrial networks.

The practising engineer in this, as in other aspects of broadcast engineering, is concerned mainly with those limitations that the discriminating viewer and listener notices—or may be expected to notice in the foreseeable future—accepting that all electronic reproduction of pictures and sound in the home is based on illusion. Fortunately, the human brain is a wonderfully adaptable and flexible system that may rapidly become desensitised to persistent defects: for example American visitors to Europe are significantly more aware of the 50 Hz 'flicker' on our pictures than are those long resident here; similarly work in Switzerland in the mid-1970s using brainwave responses showed convincingly that a viewer expecting to see interference patterns on a television picture will in fact observe them at significantly lower levels than the same observer who has not been warned to look out for them.

The r.f. engineer is, and must remain, acutely conscious of the limitations of his basic resource, the radio spectrum. Not only is this an increasingly 'scarce' natural resource (though not a resource that is diminished by use) but also it is inherently an unstable medium. Even at UHF—one of the more steady regions of the spectrum—there are problems of co-channel interference caused by anomalous

propagation (in particular that arising from the varying concentrations of water vapour in the troposphere), the 'ghosting' and signal-variations caused by reflections, and some degree of signal scattering and depolarisation such as may occur in the presence of coniferous trees and other local or terrain conditions.

Many of the developments considered in this *IBA Technical Review* would not have been possible if there had not been continued developments in semiconductor technology suitable for analogue processing, including MICs (microwave integrated circuits), YIG (Yttrium Iron Garnet) devices, CCDs (charge-coupled devices), Gasfets (Gallium Arsenide field effect transistors), very high speed (about 1 GHz) logic devices, and a number of other devices and components.

Low-noise Devices

The circuit elements with which an r.f. engineer is concerned also impose limitations, and it is perhaps in this area that he has most reason to be grateful for the steady improvement in both the active and the passive components at his disposal. At the time when the decision was made to use Bands IV and V as the main television channels in the UK, it seemed unlikely that in the foreseeable future the noise factor of domestic receivers would be less than the 12–15 dB, figures achievable with the best mass-produced thermionic valves of the early 1960s. This figure has, in practice, been reduced by a factor of at least 6 dB to about 6–8 dB—while it may become possible to design consumer-type tuners with a noise factor less than 1 dB, assuming that the price in mass-production of gallium-arsenide field effect transistors ('mesfets') comes down dramatically.

The use of GaAs fets ('gasfets') at UHF might also offer useful improvement in the dynamic range and strong-signal handling capabilities of receivers; it is unfortunate, but true, that the conventional television receiver has relatively poor electromagnetic compatibility and is often unduly vulnerable to strong out-of-band local transmissions; a factor that may prove of increasing embarrassment with the establishment of legal 'citizens' band' radio in the UK.

Oscillator Noise Sidebands

Receiver and transmitter designers are able today to take advantage of modern low-cost techniques of frequency-synthesis, with consequent benefit in stability and flexibility. However, synthesisers, combined with other improvements, have brought into much greater prominence than in past decades the question of the near-in phase and amplitude 'noise' and 'jitter'. Unfortunately the spectral purity of a phase-locked synthesised source tends to be significantly degraded from that of the best crystal oscillators—and this is a problem that is receiving increasing attention; a later section of this *Technical Review* is concerned specifically with an IBA study of the effect of the phase-noise of frequency synthesisers in broadcasting systems.

It is now almost 30 years since W. A. Edson published the first detailed study of oscillator noise (Vacuum Tube Oscillators, 1953). Edson wrote: 'It is well known that the small voltages within solid conductors and the corresponding random emission of electrons within vacuum tubes set a lower limit on the magnitude of electrical signals which may be amplified and detected... It is not so commonly realised that noise voltages also affect the operation of oscillators. It is true that in most oscillator applications the effects of noise are quite small; but in some cases, for example in microwave oscillators used in superheterodyne receivers, the noise sidebands seriously restrict the choice of IF.'

In the intervening years, as the noise contribution of solid-state amplifiers has been reduced, the practical significance of oscillator noise has increased: communications engineers for example have come to accept 'reciprocal mixing' as an important limitation on the rejection of strong adjacent channel signals.

In practice the output power from an oscillator spreads out over a band of frequencies in the form of curve associated with high-Q tuned circuits; at UHF a local oscillator may have significant noise output on both signal and 'image' frequencies, so limiting usable sensitivity. Even at lower frequencies (h.f.) oscillator noise fed into a mixer results in the receiver having some (though much reduced) response to signals over a range of frequencies corresponding to the spread of oscillator noise power. AM noise tends to fall off more slowly than FM noise; but close into the carrier the FM noise often imposes the more significant performance limitations.

In general terms, the oscillator noise performance of bipolar transistors tends to be worse than for valve oscillators; moreover the problem has been highlighted by the growing use of frequency-synthesis.

Oscillator Noise Spectrum

An oscillator can be considered as a selective amplifier with positive feedback. Noise voltages at the centre frequency build up into a continuous sine wave, until

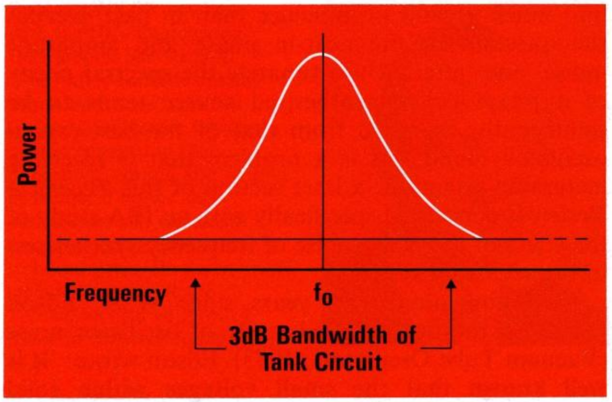


Fig. 1. Noise sidebands around a carrier (*fo*). At the frequency corresponding to the 3 dB points of the tank circuit, the noise spectrum begins to rise out of the noise plateau (−174 dBm/Hz degraded by the amplifier noise figure) at the rate of 6 dB/octave, i.e. the noise power per hertz of bandwidth increases by four times each time the offset from the carrier is halved. For a single-stage oscillator, the noise power in a 1 Hz band at a frequency δf from *fo* is equivalent to $174 - NF + Pin(\text{dBm}) + 20 \log 2Q\delta f/fo$ dB/Hz, provided $2\delta f$ is less than the 3 dB bandwidth of the tank circuit, and $174 - NF + Pin(\text{dBm})$ outside this region. *Pin* represents the power driving the base-emitter of the oscillator in dB with respect to 1 mW. A carrier plus one sideband is mathematically equivalent to simultaneous amplitude and angle modulation.

limiting action causes the oscillation to reach a steady level. At the same time noise voltages close into the centre frequency will also build up to form a roughly triangular spectrum of noise, rising above the basic flat noise output of the amplifier: Fig. 1. The width of the spectrum depends on the loaded Q of the tuned circuit and the level of the noise on the operating conditions, etc. of the circuit.

The minimum level of noise at the input of any amplifier is KTB watts, or −174 dB with respect to 1 mW/Hz of bandwidth and this will be degraded by the amplifier noise figure to set the noise plateau. If the amplifier becomes substantially non-linear due to overdriving the oscillator will be modulated by low-frequency flicker noise and there will be inter-modulation of noise sidebands. A carrier plus sideband is equivalent to simultaneous amplitude and frequency modulation of the carrier; any subsequent frequency-multiplication of the output of an oscillator will increase the FM content. To minimise the practical effects of oscillator noise, careful oscillator design is required, particularly where an output is required at UHF from an oscillator with a low fundamental frequency. Quartz oscillators can take advantage of the high Q of piezo-electric quartz but care is still required to ensure low loading of the crystal. More complex multi-stage sources, including frequency-synthesis, are likely to have degraded noise performance compared with a good crystal oscillator even when great care is taken in the design.

Measurements

It is a well-known truism that the performance of a high-grade measuring instrument should ideally be an order of magnitude better than that of the unit under test. However, with the ever-increasing technical standards required of broadcast equipment, it is clearly difficult to maintain such a differential.

An important aspect of the work of the IBA's RF Laboratory has therefore been the development of very high-performance test transmitters, receivers and precision demodulators capable of being used as practical field engineering instruments to maintain the large networks of UHF television transmitters having output powers ranging from about 0.0005–40 kW, in Channels 21–68 (470–854 MHz). To enable instruments to be used over this wide range of frequencies, it is clearly necessary to use frequency-synthesisers having very good spectral purity, and this work has thus involved investigation into the question of oscillator noise.

Synchronous Demodulation

The fundamental limitations of simple envelope demodulation of vestigial sideband transmissions have long been recognised, but it is only in the past decade that it has been practicable to make widespread use of synchronous demodulation.

Synchronous demodulation is essentially a linear frequency conversion process: Fig. 2. The modulated r.f. signal is heterodyned by a signal at the precise frequency (and phase) of the original carrier and the resulting output signal passed through a low-pass filter to remove the unwanted r.f. components, the modulation products thus being converted back to their original baseband frequencies. Whereas for an

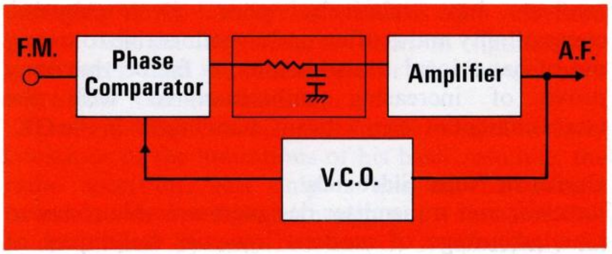


Fig. 2. The basic form of a phase-locked-loop frequency-modulation demodulator, a form of synchronous demodulation that has been facilitated by the introduction of suitable integrated circuits.

amplitude modulated signal the signal/noise performance for an envelope detector falls off markedly at low input signal/noise ratios, the synchronous demodulator preserves the signal/noise ratio, minimising signal distortion and making possible effective post-demodulator signal processing. By the use of quadrature phasing networks a PLL demodulator becomes a more flexible arrangement, providing single-sideband demodulation: Fig. 3.

The advantages of synchronous demodulation have long been known and recognised but it was not until the complexities of effectively phase-locking a local oscillator with an incoming carrier were reduced first by discrete solid-state technology and then further by integrated circuits, that this very flexible form of demodulation—which can be applied both to amplitude-modulated signals (including suppressed or reduced carrier modes) and frequency-modulated signals—has come into general use. It is now being applied widely in receivers and for instrumentation; both for demodulation of signals at a fixed intermediate frequency and also at MF/HF/VHF in the form of 'direct-conversion' or 'synchrodyne' receivers. Because the synchronous demodulator can be effective for signals at very low input signal/noise ratios, in the presence of strong adjacent-channel signals, here again the questions of oscillator spectral purity and jitter have become more rather than less important.

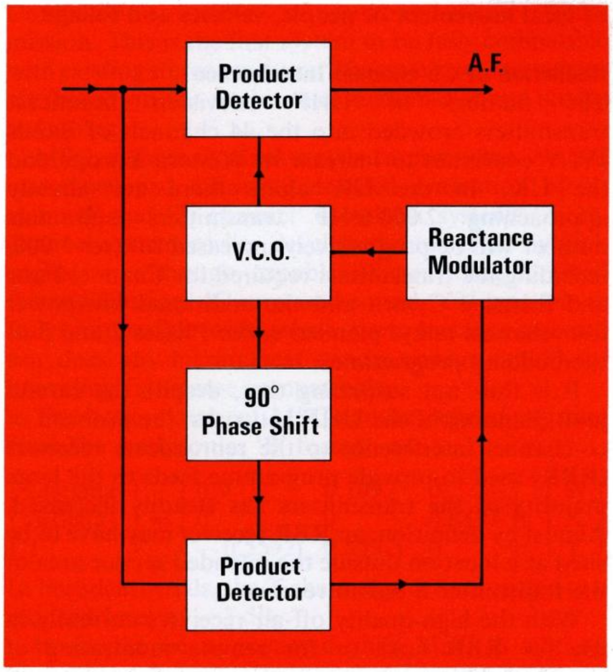


Fig. 3. The synchronous lock-loop two-path demodulator.

Aerial Arrays

The aerial array as a complex circuit element remains of fundamental importance in UHF broadcasting. The principles and practice of UHF transmitting aerials, however, have changed relatively little since the important work in the 1960s on four-channel aerials and high-power combiner units. In the United States considerable attention has been given in recent years to the question of mixed (circular) polarisation for television (as well as VHF/FM radio) but it is highly unlikely that this approach will ever be adopted in Bands IV–V in the UK where polarisation-discrimination plays an important part in permitting so many transmitters to operate within the 44 available channels.

However, the broadcaster is concerned also with high-performance receiving arrays since much use is made in the UHF network of RBR distribution: all but a small minority of the IBA transmitting stations receive programmes off-air, usually from one of the key 51 'main' transmitting stations, but occasionally as part of a 'chain' of low-power local relay transmitters. Such RBR distribution is essential from an economic viewpoint but clearly calls for the receiving equipment, including the aerial, to be of the highest possible grade in order to minimise the degradation of the technical quality of programme material. The broadcaster is, of course, vitally concerned that the viewer should lose as little as possible of the technical quality of the pictures when seen in the home, even if he has come to accept that cost factors preclude the vast majority of viewers enjoying 625-line pictures of the same standard as those seen on the monitors in the studios and the IBA's Regional Operations Centres.

Since the start of ITV colour in 1969 there has been steady improvement in the design of some of the better of the mass-produced receiving aerials, in terms of directivity, gain and bandwidth—though it is debatable as to the extent that the performance of older installations has gradually deteriorated over the years, without the viewer taking remedial action. This may be due, for example, to the effects of moisture seeping into co-axial feeder cables and/or corrosion of the aerial elements etc. There are also today many viewers taking advantage of the increased sensitivity of modern receivers to dispense altogether with the desirable well-sited, roof-top array which remains much less susceptible than set-top aerials to the effects

of local movement of people, vehicles and foliage.

Reduction of Co-channel Interference

The number of UHF television broadcast transmitters crowded into the 44 channels of Bands IV, V continues to increase in Western Europe and the UK. In the UK alone there are already approaching 2 000 UHF transmitters and this number will be progressively increased to over 3 000, including the transmitters required for Channel Four and Sianel 4 Cymru and the additional low-power four-channel relays planned under Phases 2 and 3 of the building programme.

It is thus not surprising that, despite the careful joint-planning of the UHF networks, the problem of co-channel interference to the rebroadcast receivers (RBR) used to provide programme feeds to the large majority of the transmitters has steadily increased. Almost by definition, an RBR receiver may have to be sited at a location outside the intended service area of the transmitter it has to receive.

With the high-quality off-air receivers currently in use for RBR links or for remote monitoring of unattended transmitters, the quality of the received signal is determined primarily by the propagation path (in terms of signal levels, fading and multipath) and by the level of co-channel interference (CCI) at the receiving site, which again will be affected by the propagation path(s) between the receiving site and the unwanted transmitter(s).

The provision of suitable sites and propagation paths for the wanted signals of RBR links is basically a function of the service planning engineer. Limitations on choice of site and topography may require the use of high-gain aerials, mounted well above ground level, to provide sufficient gain; on some difficult paths, particularly those involving sea paths, some form of diversity reception may be desirable to minimise fading.

For many years, the reduction of CCI has been viewed primarily as a question of aerial directivity, designing or adapting a high-gain array so that a null point of minimum pick up exists in the direction of the interfering station. This technique does, in practice, usually cope adequately with CCI from one source, or even a few sources, provided that none has a bearing very close to that of the wanted transmission: it is also, of course, necessary to be in a position where the design engineer can take into account future likely sources of interference.

The growth in the number of existing or potential sources of interference combined with the variable 'anomalous' propagation characteristics of UHF signals has made the elimination of CCI at some sites steadily more difficult. Fixed arrays providing nulls of more than 30 dB are difficult to implement as they impose demanding requirements on pointing accuracy and rigidity of a high structure; nor can they cope with any change in the direction of arrival of incoming signals.

Since the early 1970s, the investigation of CCI and methods of combating it have been under almost continuous investigation by the IBA's Experimental and Development Department and in particular by RF Section. From the start it was recognised that the most promising solutions to this problem—at least in situations where any solution was feasible—were to be found in the use of adaptive systems that would adjust automatically to changing conditions: work has therefore been concentrated on adaptive aerial systems of varying complexity, and on potentially rather simpler adaptive filtering at intermediate frequency. This work, including the highly successful four-channel SABRE (Steerable Adaptive Broadcast Receiving Equipment) array erected on Alderney, is described in detail in this volume—although it should be appreciated that SABRE is a highly sophisticated and complex system that is unlikely to be widely duplicated. It meets the particularly difficult situation of receiving high-quality UHF colour transmissions over a sea path exceeding 100 km in an area where an unusually large number of potential sources of interference can be expected, and where tropospheric propagation effects are of above-average occurrence.

Where all potential sources use 'carrier offset' frequencies the possibility arises of using adaptive transversal filtering at intermediate or even video baseband frequencies, without the necessity of using an adaptive aerial array. Since such a filter cannot reject interference from a non-offset transmitter, and inevitably introduces some degree of waveform degradation, it would not have provided a complete solution to the CCI problem existing at Alderney.

The SABRE adaptive array at Alderney is a four-channel system using digital logic to vary the phase and amplitude of signals from the individual aerial elements, providing a receiving array that can automatically and continuously adjust its polar diagram to minimise interference from up to twelve unwanted transmitters by providing nulls of up to 45 dB depth in the reception pattern while simultaneously receiving wanted signals on up to four programme channels arriving from a co-sited UHF mainland transmitter.

It was the first successful use of a fully adaptive array for this type of application; unlike previous communications and radar arrays it was designed for optimising signal-to-noise ratios, rather than optimising signal gain. The use of electronically-steered complex phased arrays has had a chequered history, including experimental h.f. arrays ('Musa') in the USA; the 'Medusa' h.f. array built on Cooling Marshes, Kent by British Telecom (formerly BPO); and the large Wullenweber direction-finding arrays developed initially by the Germans during World War II. More recently the development of electronically-steerable radar arrays included a truly enormous structure in the USA that was destroyed accidentally before it was commissioned.

The first investigation into the use of filtering techniques for CCI suppression was carried out by IBA engineers as early as 1971. This consisted of a simple carrier-suppression system for possible use in off-air monitoring. In 1973 a transversal filter using delay lines was investigated for RBR applications and an experimental version built. Adaptive filtering has been similarly investigated although the system is not an operational installation at this stage.

An adaptive aerial system, though more complex, provides a form of spatial filter (i.e. rejects signals according to direction). The interfering signal can be totally rejected to the limit of the null depth, including chrominance components, with negligible effect on the wanted video signal.

An adaptive filter system, essentially a form of comb filter, can reject luminance components close to the line frequency, including harmonics, from up to two sources of offset CCI, and partially reject such components from other sources. It cannot reject any chrominance components of unwanted signals. This means that the subjective improvement is limited by the vertical rate of change of the picture information content of the unwanted signals. These, and some other effects, limit the subjective improvement of co-channel protection ratio to about 12 dB. However, such a filter is, of course, considerably less complex and less costly than even a simple adaptive aerial system. The two approaches are thus seen as complementary and capable of providing powerful methods of improving the quality of signals received off-air.

Improved Demultiplexer

An earlier volume of *IBA Technical Review* has described the work that led to the development by IBA engineers of the MSC ('Mono Stereo Compatible') system of surround-sound transmission. This is the first system to be fully compatible with existing stereo receivers. The system was tested on-air (based on the Ambisonics surround-sound studio technology).

MSC, a three-channel matrix transmission system, was developed following detailed evaluation of '2-channel' and '2½-channel' broadcast systems. A 2½-channel system earlier developed and tested by IBA engineers, although capable of providing a satisfactory illusion of surround-sound, did not prove fully compatible with stereo reproduction; nor did nor does any hierarchical system appear capable of providing full stereo compatibility.

The basic matrix of MSC can be defined, starting from the Ambisonics 'B' format, as:

$$\begin{bmatrix} \Sigma(\text{sigma}) \\ \Delta(\text{delta}) \\ \text{T} \end{bmatrix} = \begin{bmatrix} 0.9 & 0.1092 & 0 \\ 0 & 0 & 0.6897 \\ -0.5592\text{j} & 0.5592\text{j} & 0 \end{bmatrix} \begin{bmatrix} \text{W} \\ \text{X} \\ \text{Y} \end{bmatrix}$$

In broadcast trials, the T channel was transmitted at a reduced level of −3 dB.

It should be appreciated that MSC is not an hierarchical (universal) system and optimum results cannot be achieved by listeners using a Matrix H, HJ or UHJ decoder.

During 1975–80 the IBA team investigated surround-sound, theoretically and practically, along the entire transmission chain from microphones to loudspeakers. This work convinced the team that to be successful surround-sound must provide the highest achievable technical quality and offer a significant improvement over existing broadcast stereo, while not in any way degrading quality for the majority of listeners who, for many years, would inevitably continue using conventional mono or stereo equipment. It was also fully recognised that any international standard for surround-sound must also be suitable for the recording industry, particularly in conjunction with the new digital-audio recording systems.

However, by the end of 1980 it was becoming apparent that it would take many years to achieve international agreement on a transmission system for surround-sound; there was also the further problem that at an early stage the BBC had stated that all its surround-sound transmissions would be receivable on Matrix H decoders.

For these reasons, the IBA have no present plans to provide surround-sound on ILR, although it is hopeful that in time the merits of the MSC system will be recognised nationally and internationally, and that

it will be possible to resume work on the system. In the meantime, the improved demultiplexer, initially developed for MSC, is equally valuable for stereo and/or three-channel transmission systems regardless of the form of matrix. An important consideration is that the decoder provides much improved adjacent channel rejection compared with conventional decoders.

PROPAGATION ANOMALIES

There is little that an r.f. engineer can do to change the inherent unwanted anomalies of VHF/UHF propagation. The challenge is to find circuit and aerial elements that will minimise the harmful effects. An early example was the introduction in the early 1930s of automatic gain control (AGC) in radio receivers to combat fading; unfortunately, a simple feedback loop can do little to overcome the problem of frequency selective fading—though this can be diminished with such modes as single sideband transmission.

Tropospheric Propagation

To the broadcast engineer trans-horizon propagation of VHF, UHF and SHF signals is mostly an unwanted anomaly, a major cause of co-channel interference. While it is usual to think of such trans-horizon propagation as rare, it exists to some significant degree, particularly over sea paths, for a very significant percentage of the total time; this can amount in some circumstances, in some parts of the UHF spectrum, to as much as 10% of the total time.

Extensive investigations have been made by both broadcast and communications engineers of the extent of trans-horizon propagation and the weather conditions that give rise to it. The following notes, which provide useful background information to the work aimed at reducing CCI, are based on investigations carried out at 11 GHz by M. T. Hewitt and A. R. Adams of British Telecom Research as reported to a URSI Symposium in 1980.

Figure 4 shows the main features of the Earth's atmosphere and structure affecting radio propagation at all frequencies. For UHF terrestrial networks, television engineers are interested primarily in the lower atmosphere (Fig. 5). However, direct-broadcast satellites involve SHF propagation through all ionospheric layers.

Enhanced propagation is most frequently brought about by the action of water vapour in the region of a mile or so above the Earth. For sea paths around the southern and eastern coasts of the UK the most consistent time for super-refraction conditions is about 1800 GMT and such propagation tends to be minimum about 0800 GMT. The main weather patterns that result in super-refractive layers in the lower atmosphere, as found by Hewitt and Adams, are as follows:

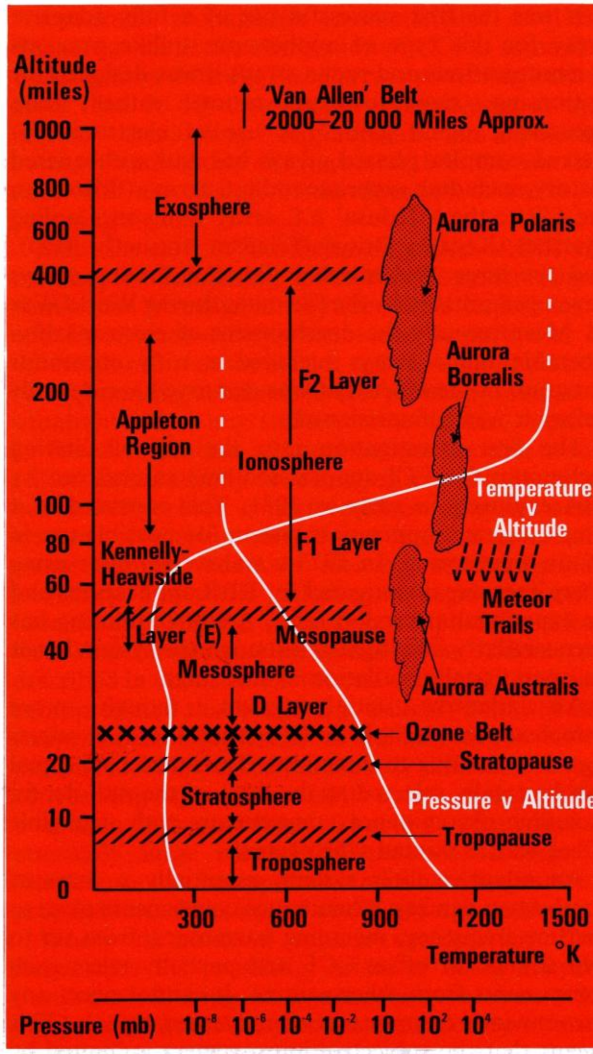


Fig. 4. The Earth's atmosphere and structure showing the main layers that can affect radio propagation. The diagram also shows the normal temperature versus altitude and pressure versus altitude readings. A reversal of the normal decrease of temperature with height in the troposphere will affect the distribution of water vapour and is likely to result in enhanced tropospheric propagation and ducting. (Source 'VHF radio wave propagation' by W. G. Burrows).

Subsidence

Air which slowly descends within an anticyclonic

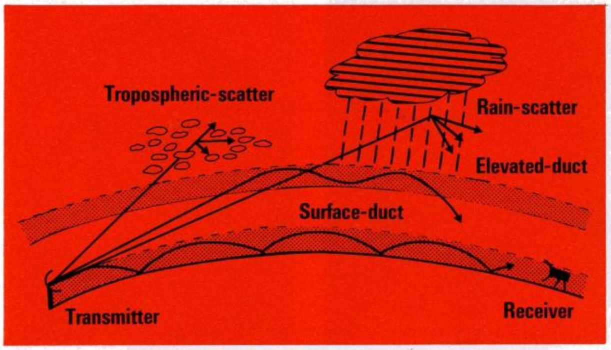


Fig. 5. The main trans-horizon propagation mechanisms in the lower atmosphere affecting (in varying degrees) UHF and SHF signals. Signals propagated by either surface or elevated ducts may be sufficiently strong to affect reception on domestic receivers or RBR links. Auroral reflections may also occasionally cause interference. Rain scatter becomes of significance at frequencies above about 10 GHz but has negligible influence on UHF television propagation. Based on the work of Hewitt and Adams, British Telecom Research.

system becomes progressively heated by compression; this subsidence is halted when opposed by forces in the air below. The warmer air then spreads out in a layer without mixing with the lower, cooler air. When the subsidence inversion is about 2 000 m above the surface, it imparts stability to the lower air, suppressing convection currents. Where the subsidence inversion descends still lower, there will be a very pronounced inversion layer which may last several days. The lower the layer the stronger the trans-horizon signals will tend to be. Subsidence inversions can be identified meteorologically from radiosonde data; this form of inversion can exist over land or sea.

Advection

The air flow around anticyclonic systems over central Europe produces a drift of warm dry air out over the cooler very moist air just above the sea. When, in the absence of winds, there is little convection turbulence, this overlay of warm air continues without mixing. It is this form of layer that is the main reason for the enhancement of SHF signals for more than 10% of the total time, although it would appear that the signal enhancements are seldom as pronounced as those of very low subsidence inversions. Advection inversions occur primarily over the sea but also can occur in low-lying coastal areas. They can be identified from meteorological data by differences in temperature and humidity, though problems can arise in linking data with inversions.

Radiation Cooling

After a warm day and a calm, cloudless night, there will often be considerable cooling of land surfaces, and surface ducts may be formed. In fact radiation nights are the most frequent cause of signal enhancement over land areas in the UK, but the events tend to be of short duration. Day-night temperature difference, cloud cover and wind speed can all provide reliable indications of radiation cooling. For a given degree of surface cooling there is likely to be a trade-off between height and the refractivity lapse rate, so that signal enhancements will be less and fewer as the 'roughness' of the intervening path increases (i.e. hills etc.).

Weather Fronts

Some signal enhancements are due to the movement of weather fronts into an area of a weakening anticyclonic system. Such relatively short periods of trans-horizon propagation result from super-refractivity somewhere within the frontal structure, but further investigation is needed as these frontal disturbances are the most difficult of all the mechanisms to detect from meteorological data. No reliable indicators have yet been found and this remains an important area of work to be completed.

Multipath Distortion

Multipath propagation—the arrival of 'echo' signals delayed in time from the main signals—is an important source of degradation at HF, VHF and UHF. Whereas at HF the problem tends to arise from reflections from different layers and 'patches' of the ionosphere, on VHF and UHF the usual source of reflection is a physical structure: hills and mountains, tall buildings, gas holders, towers, masts, cranes and trees.

The strength of the echoes may vary seasonally or over a period of time due to changes in foliage, new building work and the like. The delay or 'term' of the echo depends on the difference in path length of the direct and indirect signals: short-term echoes can affect teletext data reception and chrominance/luminance ratios but have little noticeable effect on VHF/FM sound. Long-term echoes (typically delay times of two microseconds or more) produce visible 'ghosting' on television pictures and distortion on sustained high-frequency notes on VHF/FM radio; another effect is to reduce channel separation of multiplexed stereo or SCA channels. Distortion products are generated at many frequencies, not only those harmonically related: Fig. 6.

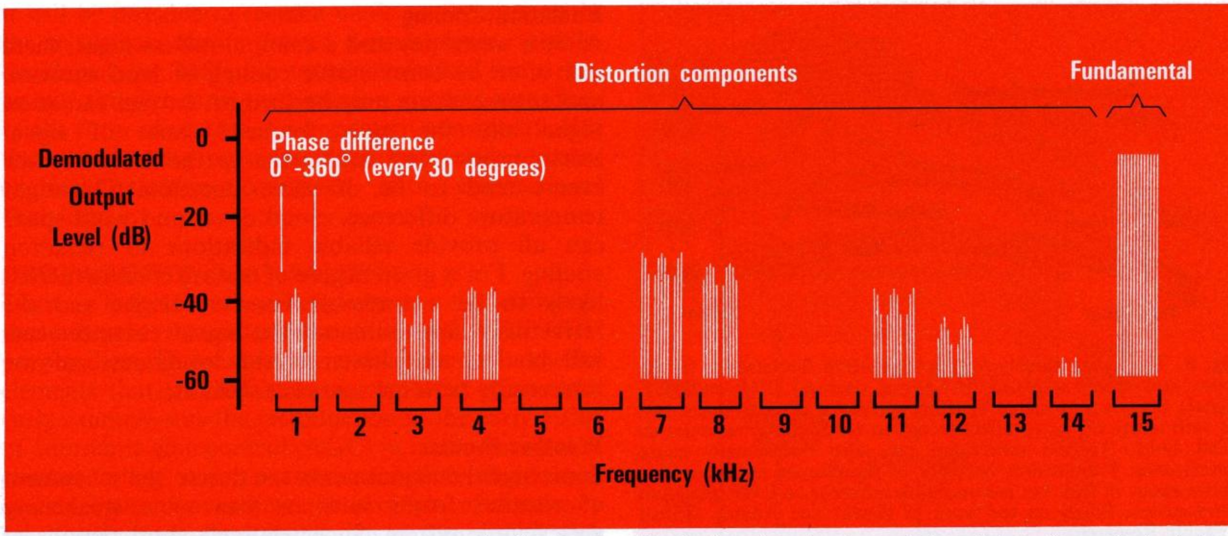


Fig. 6. An NHK computed example of the spectrum distribution of multipath distortion, showing variation of distortion products at phase differences 0°–360° (in 30° steps) with a desired/undesired signal ratio of 20 dB and with an indirect path delay of 10 μs. Modulation of left-hand signal 15 kHz, 100%. The maximum distortion is 5.6% with a phase difference of 180°. Minimum distortion 4% (with phase differences of 90° or 270°). Note that distortion components are produced at frequencies not harmonically related to the fundamental signals.

Recent work by NHK engineers has underlined that multipath is one of the major factors that degrade the reception of high quality radio both on domestic and car-radio receivers, and is significantly more severe on stereo than on mono transmissions:

(*a*) Distortion tends to be more pronounced if the delay time of the reflected unwanted signal with respect to the direct wanted signal is comparatively long (Fig. 7).
(*b*) The distortion is almost inversely proportional to the D/U (desired/undesired) ratio if this ratio exceeds 10 dB.
(*c*) The phase difference which gives maximum or minimum distortion is not constant but varies with such parameters as delay time, modulation frequency and depth of modulation.
(*d*) Maximum distortion at 15 kHz is greater than at lower audio frequencies, and is several times greater than at 1 kHz.

VHF/FM receivers in which there is a poor standard of AM rejection produce significantly more distortion than where this is not the case. But it must be recognised that all receivers suffer from multipath distortion and that at present the only practical way of minimising this is by increasing the D/U ratio of the signal provided by the aerial. One method of doing this would be by using phasing and adaptive techniques to improve aerial directivity and to null out the unwanted delayed transmission. For television 'ghosting' with AM signals it is possible to use electronic delay lines such as CCD units to provide phase-cancellation of the unwanted signals.

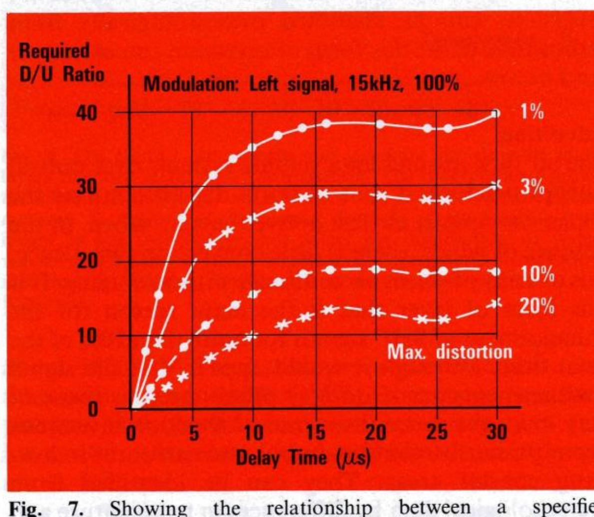
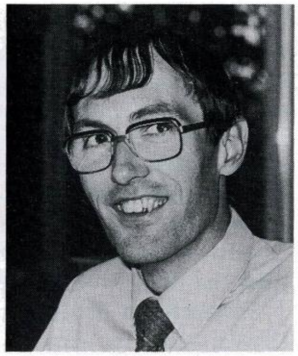
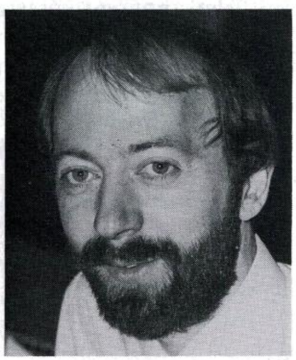


Fig. 7. Showing the relationship between a specified desired/undesired signal ratio and delay time (path difference between the desired and undesired signals) for various degrees of multipath distortion as calculated by NHK.

MICHAEL WINDRAM, MA, Ph.D., C.Eng., MIEE. A Biographical note on Dr Windram appears on page 41.



LLOYD BRUNT, M.Sc., C.Eng., MIEE, graduated from Portsmouth Polytechnic in 1969 and, with industrial sponsorship from the Plessey Company, completed a post-graduate degree at Birmingham University in 1971. He joined the IBA Radio Frequency Section of The Experimental & Development Department in 1975 and is currently a Senior Engineer there.



EDGAR WILSON, B.Sc.(Eng.), ACGI, studied at Imperial College, London following a pre-university year with BBC Designs Department. After graduating in 1972 he joined the British Post Office (now British Telecom) to work on digital telecommunications systems. He joined the IBA's Experimental and Development Department in 1974. As a member of RF Section he worked on the SABRE project and subsequently, with Video and Colour Section, on the development of a 60 Mbit/s digital codec for satellite TV. He is now a principal engineer in RF Section.

Adaptive Aerials for UHF RBR Links

by M. D. Windram, L. Brunt and E. J. Wilson

Synopsis
The large and still increasing number of UHF television transmitters operating in Bands IV and V has brought about an increased risk of co-channel interference (CCI) in the rebroadcast receiver (RBR) links that are used to provide programme feeds for the vast majority of transmitters. Particularly formidable problems faced those planning a high-grade RBR link over the 135 km sea-path between Stockland Hill, Devon and Alderney, the nearest of the Channel Islands. To provide a solution the IBA engineers developed an adaptive aerial array which automatically and continuously adjusts its polar diagram and in which the constraint is that of minimising the level of CCI relative to the wanted signal. This paper discusses the basic requirements and design of complex adaptive arrays and the implementation of a system constructed as a 2×2 array of 8×2 dipoles. The array provides a gain of about 24 dB and has a continuously-adjustable rejection performance of about 45 dB or some 20 dB better than can normally be achieved using even a complex fixed array. The array has to reject unwanted signals only 7° from the bearing of the wanted signal. Some suggestions are made for simpler four-element adaptive arrays which would be suitable for difficult sites, including island sites, where there are fewer potential sources of interference.

By early 1982, the total number of transmitters operated by the Independent Broadcasting Authority in the UHF band (Bands IV and V) was over 700. It is expected that, by completion of the UHF network for the existing ITV channel, the number will have risen to about 1,000. This means that when the fourth television channel is fully engineered the total number of UHF transmitters in operation in the United Kingdom may well be around 4,000, crowded into a total of 44 available channels in the UHF band. Although the distribution of transmitters is being carefully planned to avoid interference problems within the appropriate programme service areas, effects such as anomalous propagation cause

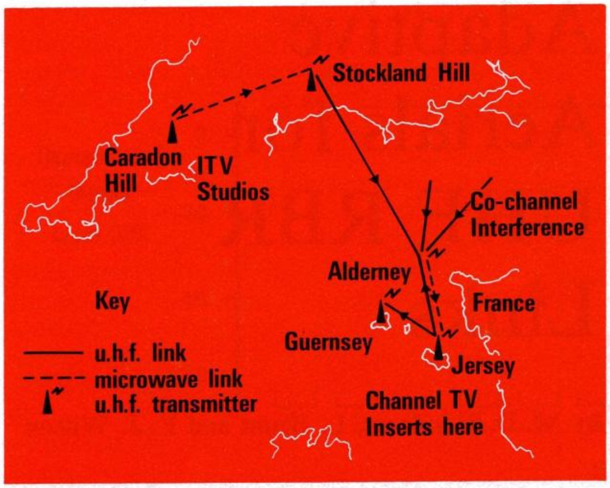


Fig. 1. In 1974, following investigation of a number of alternative proposals for providing a high-quality 625-line colour programme link between the UK and the Channel Islands, it was concluded that the most satisfactory solution would be to re-engineer the existing 405-line link receiving station on Alderney, providing a new aerial designed to receive the UHF broadcast transmissions from the 250 kW transmitting station at Stockland Hill, near Honiton, South Devon and then to provide a programme feed to Jersey by means of a special microwave link. While signal strength of the Stockland Hill transmitters was considered just adequate (provided space diversity aerials were used) it was recognised that there were many potential sources of co-channel interference.

considerable problems to broadcasters who need to pick up signals outside these service areas as part of the broadcasting network.

The UHF link from Stockland Hill, Devon, to Alderney, which forms part of the broadcast network for the Channel Islands, as shown in Fig. 1, is of particular interest. This is perhaps the most difficult UHF link in the broadcasting network and many alternative routes were considered before this was finally chosen. It is for this link that the first RBR adaptive aerial system was developed. There are, however, potentially several other UHF links in the United Kingdom which, although not exhibiting quite such severe propagation characteristics, are nevertheless still a problem and for which a simplified form of adaptive aerial system would be an ideal solution.

Requirements and Design of Adaptive Array

The UHF path from Stockland Hill to Alderney is an over-the-horizon sea path of some 135 km in length, and, characteristic of such paths, the received signal is very variable in strength with a range ~60 dB and generally very weak. For this reason, the signal is susceptible to co-channel interference (CCI), from both existing and proposed transmitters as shown in Fig. 2. Kippure for example is at a particularly narrow angle off the wanted signal of 7°. Figure 3 shows the bearing and field strengths of the interference. The inner ends of the lines represent the median field strengths and the outer ends represent the field strengths exceeded for 1% of the time. Note for example that the field strength of Crystal Palace, London is for 1% of the time more than 6 dB greater than the median field strength of Stockland Hill.

To obtain a broadcast quality signal, it has been shown, both by theoretical propagation predictions and by practical measurements, that the reception pattern of an aerial on Alderney needs on occasions to have null depths of the order of 45 dB in the directions of the interfering sources. It is not possible to use conventional aerials for this degree of rejection for the following reasons:

(*a*) Initial mechanical and electrical assembly tolerances of conventional aerials tend to limit the designed null depths to 25–30 dB.

(*b*) The pointing accuracy required for very deep nulls is extremely high, setting impossibly high constraints on mounting tolerances.

(*c*) The apparent direction of the interference can vary with time. Such variations may be caused by combinations of propagation mechanisms, such as tropospheric scatter etc.

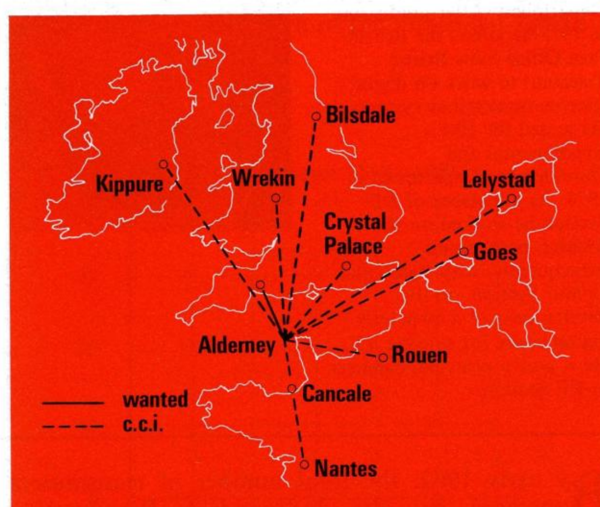


Fig. 2. IBA propagation studies showed that harmful interference levels from a number of existing or planned transmitters would exist for more than the acceptable 1% of the time. Some but not all transmitters would use 'offset' carrier frequencies, and some interference might be expected from stations such as Kippure (Ireland) along paths close to the wanted signals from Stockland Hill.

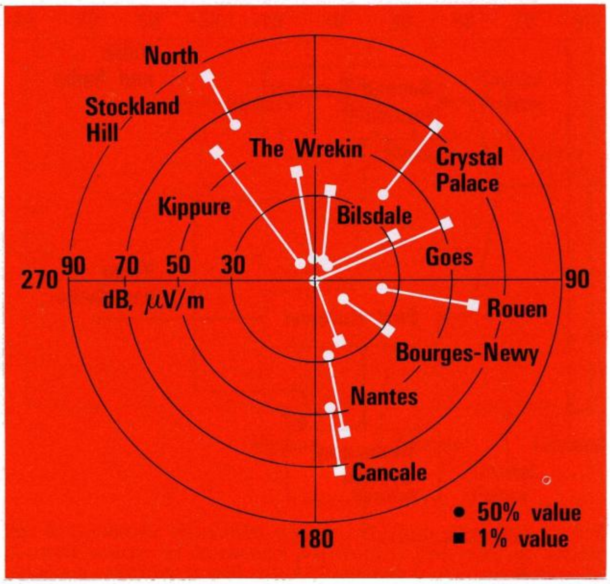


Fig. 3. A polar plot of predicted field strengths against bearings of potential sources of CCI. The inner (weaker) signals represent the median (50%) values and the outer ends the field strengths likely to be exceeded for 1% of the time (e.g. during pronounced tropospheric propagation conditions). Considerable aerial directivity with nulls of the order of 45 dB is clearly required if CCI is to be maintained within acceptable protection ratios for 99% of the time.

For these reasons, the IBA decided to investigate the properties of adaptive arrays. The advantages of such an array include:

(i) automatic adjustment of the aerial pattern to give minimum interference.

(ii) no need for prior knowledge of the bearing of the interference; the array is therefore able to handle interference from sources not previously predicted.

(iii) ability to handle multiple sources of interference up to a limit which can be defined.

(iv) ability to track any apparent changing of direction of a source that might result from propagation effects.

(v) there is no longer a severe mounting and aerial tolerancing problem.

After a period of experimentation, including the installation of a prototype half-size system at Alderney to confirm earlier theory, the operational adaptive aerial was installed on Alderney in March 1977. The block diagram of this system is shown in Fig. 4.

The adaptive array is based on a linear array of sixteen elements, the output of each element being connected to a network which effectively controls the amplitude and phase of that output. These signals are then combined, and it is this process which creates a voltage-controlled aerial, the pattern of which is a function of the control voltages. The receiver or receivers used provide the video and audio outputs required and also provide signals for the measuring system. Up to four receivers may be used within the control loop. The adaptive control unit provides the logic which causes the aerial voltages to change in the manner required to reduce the interference.

To develop a successful adaptive array major difficulties had to be solved. These included the design of the control networks, the measuring system in conjunction with the receiver and the type of algorithm to be used. To assist in determining the requirements, a detailed theoretical study of adaptive arrays was carried out and is reproduced in part in the Appendix to this volume ('Adaptive Aerial Arrays—A Theoretical Introduction').

Conclusions from Theory

Initial analysis showed that the ideal element spacing is $\sim\frac{2}{3}$, which combines reasonable directivity of the array with ease of control of the nulls. Cartesian $(x+jy)$-type control of the aerial outputs was chosen because it avoids the serious problem of discontinuities in the controls that occurs with pure phase shifters having only a finite phase range. This is illustrated in Fig. 5.

With polar control, a change of weight from A to B involves a discontinuity, since phase has a limited range. For Cartesian control there is no discontinuity so that any small incremental step is possible, in any direction, within the control domain. This is essential in an adaptive process in which small changes of controls may be required and for which discontinuities could lead to instability. This also makes the use of the constant gain criteria easier.

For arrays of the size used at Alderney at UHF it is necessary because of expense, stability and maintainability to use control algorithms which take measurements by making step changes in the control voltages to the array, thus stepping the aerial pattern and measuring the result in terms of CCI on the output signal. Other methods such as correlation techniques are far too expensive in terms of hardware and complexity, although in principle they are capable of giving more accurate measurements of the error; they were therefore not considered in detail.

The various algorithms discussed in the Appendix were fully considered. Of these, for the Alderney

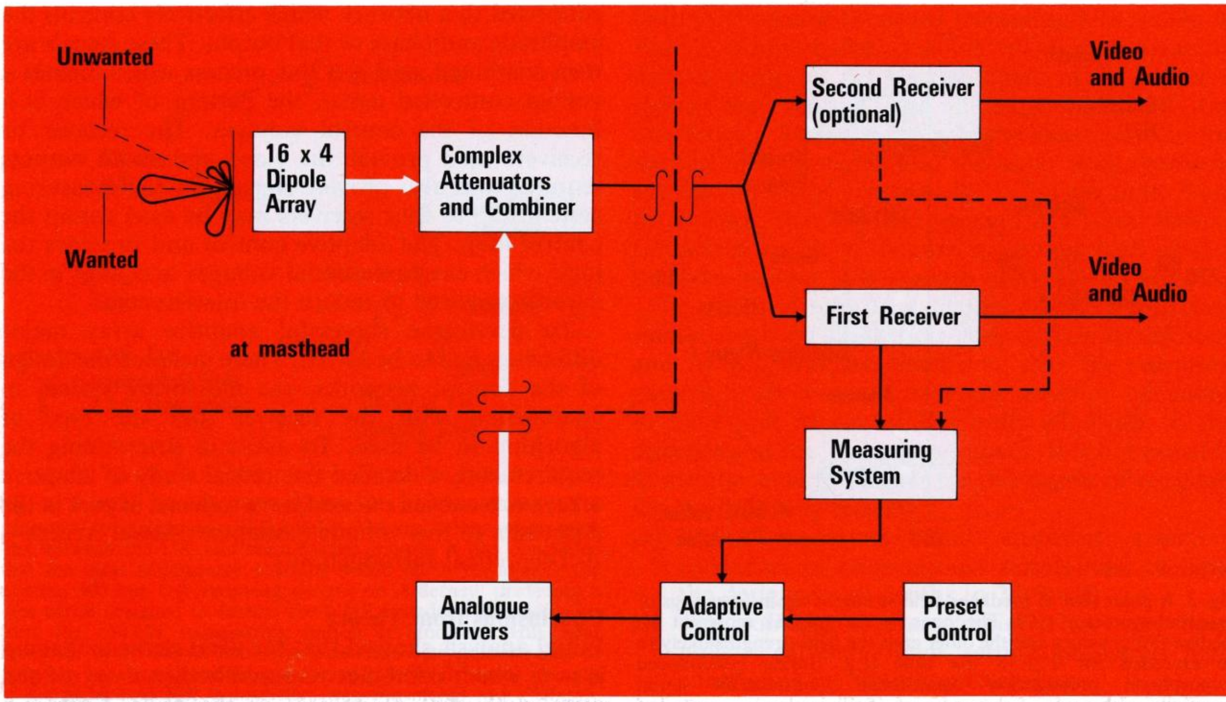


Fig. 4. Following a period of experimentation, including the installation of a prototype half-size system on Alderney to confirm the theoretical studies, a multi-channel linear array of 16 × 4 dipoles with adaptive control was designed and implemented to become the first operational adaptive system for RBR applications. Each element is connected through a network which controls the amplitude and phase of the output, providing a voltage-controlled aerial pattern. Up to four receivers may be used within the control loop, and the aerial pattern is continuously adjusted to optimise signal/noise ratios of the desired signals.

application, the simple hill-climb algorithm seemed to be most suitable for operation close to the minimum interference condition. This is the normal condition for the array, since much of the operation of the adaptive array for broadcasting purposes is involved in maintaining nulls on CCI sources rather than steering them onto sources. The simple algorithm has the advantage of being easy to implement and requires only a knowledge of the sign of a step, not the magnitude which can be very inaccurate in the presence of noise. It also does not require a large number of measurements, each taking a set time, as some of the more complex algorithms require.

Figure 6 shows the results of a computer simulation for five sources of interference at angles of $-7°$ (Kippure), $+65.4°$ (Crystal Palace), $+23.6°$ (The Wrekin), $+35.4°$ (Bilsdale) and $+85.2°$ (Lelystad). The patterns are the horizontal voltage radiation patterns of the array, starting at the top with the initial (preset) radiation pattern which is that of a uniform array, and with plots at intervals corresponding to approximately 5 secs in real time until an optimised pattern is reached. The final null depths are 50 dB for each of the four sources of CCI.

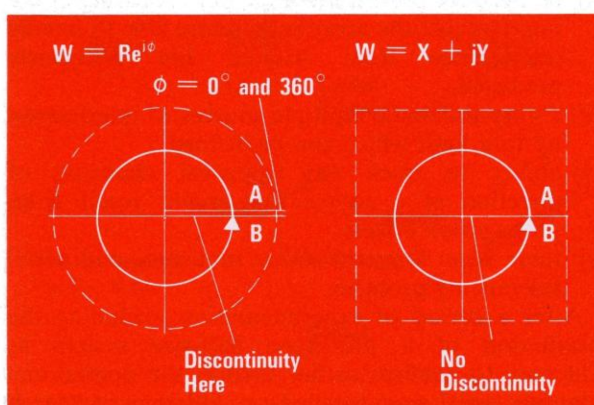


Fig. 5. Polar and cartesian control of amplitude and phase. Cartesian (X+jY) type control was chosen in preference to polar control since it avoids the problem of discontinuities in the control that would occur with pure phase shifters having only a finite range of phase control.

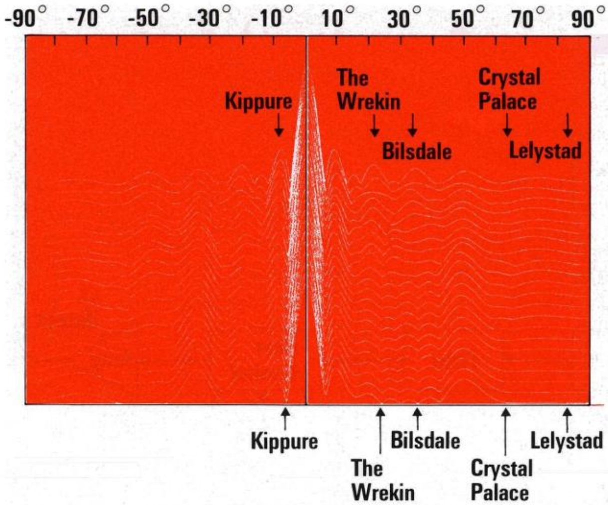


Fig. 6. Computer simulation using a hill-climb algorithm for five potential sources of co-channel interference. The relatively simple hill-climb algorithm is well suited for operation of an array close to minimum interference condition since for much of the time the prime requirement of the array is to maintain nulls on known potential sources of CCI rather than steering the nulls on to less likely sources of interference.

Operational Adaptive Array

(*a*) *Array*

For Alderney, the aerial requirements were not only the horizontal distribution of 16 elements at $\sim\frac{2}{3}\lambda$ spacing, but also a gain of ~ 24 dB. This was achieved as a 16×4 dipole array constructed as a 2×2 array of 8×2 dipoles. The 8×2 dipole array was constructed with a solid backplane and used eight printed circuit panels, each consisting of two bat-wing dipoles with printed feeders to a single point. This design combined wide bandwidth with good reproducibility, performance and ease of construction. The dipoles are protected from the direct effects of weather and salt spray by a fibreglass front cover to the box structure holding the 8×2 dipole array. The aerial array on the tower at Alderney is shown in Fig. 7, and Fig. 8 shows the arrangement of a single module with the cover removed.

The four modules are mounted as a 2×2 array at a height of ~ 23 m up the tower at Alderney. Because of the modular construction, necessitated by mechanical and structural requirements, the array is in effect set up as a 17-element array with the middle element missing. Computer simulation and actual measurements show no significant difference in behaviour from that of a 16-element array.

(*b*) *Masthead box*

The amplitude and phase adjustments, and combining, are all carried out in a box mounted behind the aerial. The block diagram is shown in Fig. 9.

The filters are required because there is a local transmitting antenna mounted on the same tower. The amplitude and phase adjustment is carried out by the complex attenuator, so called because it multiplies the signal by the complex number $x+jy$. The block

Fig. 7. The SABRE array installed on the Alderney tower (the elements are in the flat panels). The masthead box containing the complex attenuators, filters and preamplifiers is mounted on the access platform immediately behind the array.

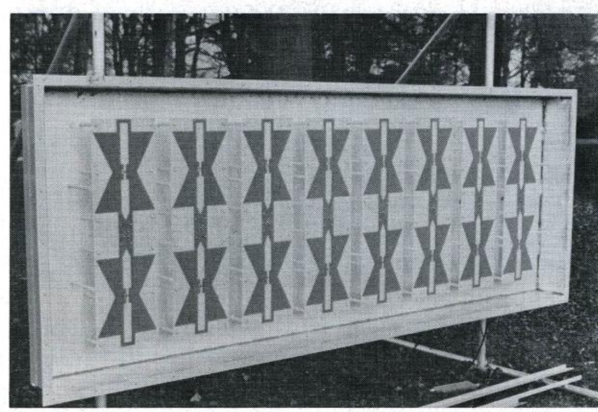


Fig. 8. An 8×2 dipole array module shown with the protective front cover removed. In the operational model, the four modules are mounted as a 2×2 array at a height above ground of approximately 23 m. To meet structural requirement, the array is in the form of a 17-element array with the middle element missing but measurements and computer simulation show no significant differences from that of a 16-element array.

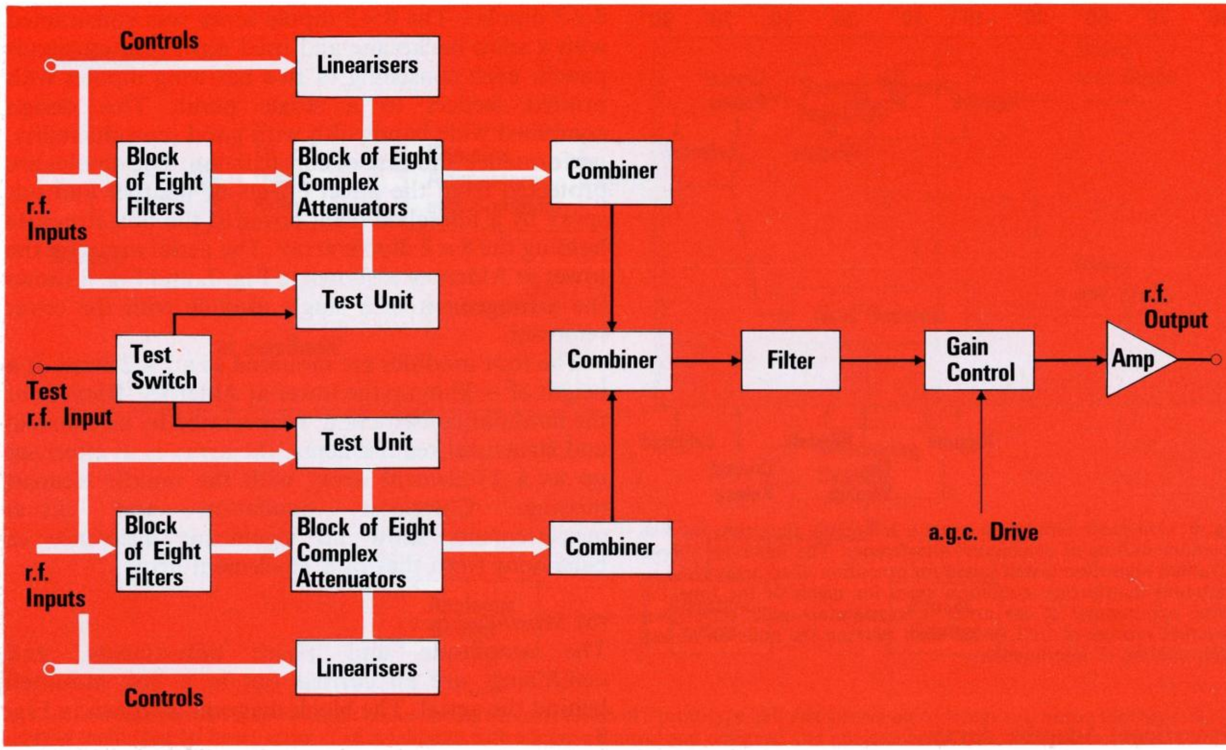


Fig. 9. Block diagram of the masthead box which carries the amplitude and phase complex attenuators and the combining units together with selective filter to overcome any electromagnetic compatibility problems arising from the presence of a local transmitting aerial mounted on the same tower.

diagram of an individual complex attenuator unit which is a single printed-circuit board is shown in Fig. 10. Figure 11 shows a photograph of the complex attenuator.

Use of the reflective type of attenuator is important in that it enables the gain in either the in-phase or quadrature path to be any value in the range -1 through zero to $+1$ for a perfect attenuator. The performance of the complex-attenuator network with appropriate linearisers has been found to be very stable and to give a transfer characteristic closely approaching the ideal. This is essential for the constant-gain algorithm to operate correctly.

Because the box is mounted at masthead, complete testing facilities are provided so that, by means of *p-i-n* diode switches, a test signal can be routed into any one of the attenuators to enable the attenuator to be checked. If the test signal is switched to a particular attenuator during normal operation, then it appears as interference and the system automatically adjusts the appropriate control voltages of the attenuator to zero; this state can readily be checked.

The masthead box also incorporates a voltage-controlled attenuator which forms an extension to the receiver AGC. The drive to this AGC is a single digital line which causes the gain to increase or decrease slowly with time. This has been designed to operate not only in a system such as the adaptive aerial, but also for masthead pre-amplifiers and can be driven from any combination of receiver AGC system via a simple comparator unit. A photograph of the masthead box is shown in Fig. 12.

(*c*) *Receiver*

From the masthead box, the signal is passed down to the receiver or receivers. The type used is a modification of a receiver designed by the IBA and currently in use at the majority of its higher-power rebroadcast relay transmitter sites. The main feature of the unmodified receiver is synchronous detection which has been successfully used by the IBA for some years now to eliminate quadrature distortion which would otherwise result from envelope detection of a vestigial-sideband filtered signal. The detector used has

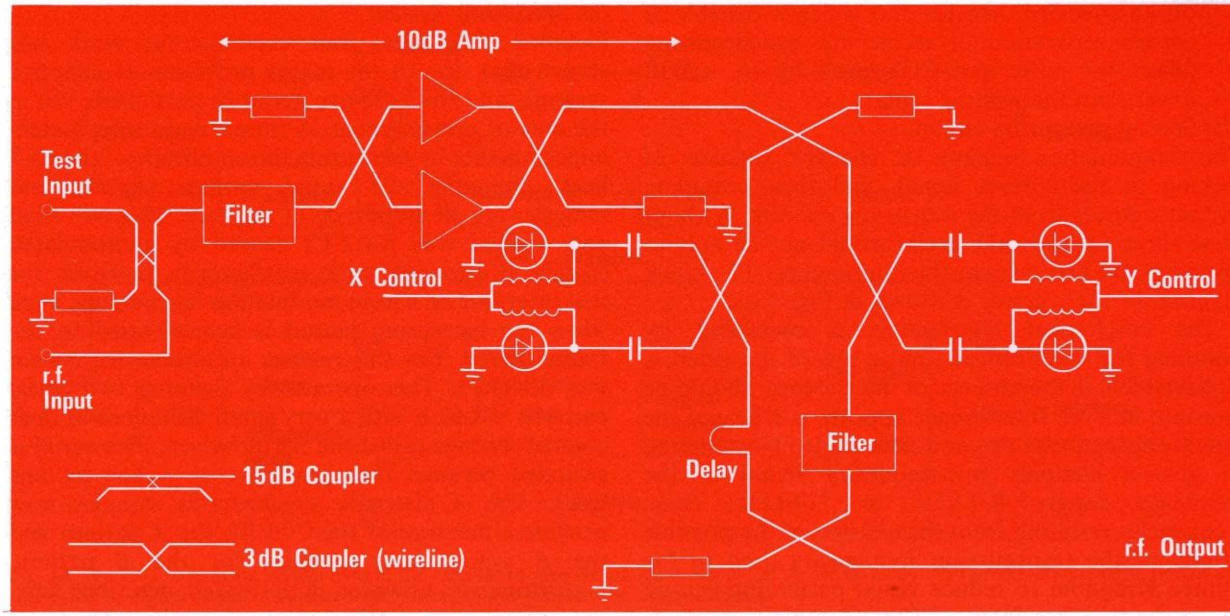


Fig. 10. Block diagram of a complex attenuator of which 16, one for each aerial element, are fitted in a masthead box. After amplification, the signal is split into two separate attenuators; subsequently the two processed signals are recombined in phase-quadrature. These complex attenuators have proved to be operationally very stable with a transfer characteristic approaching the ideal, permitting the constant-gain algorithm to function correctly.

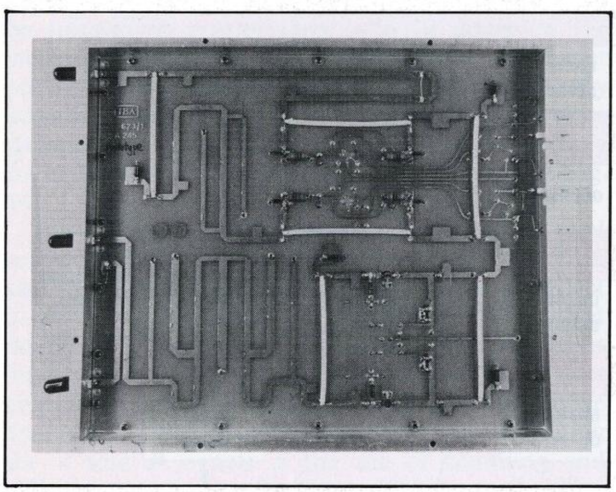


Fig. 11. A complex attenuator of the type outlined in Fig. 10. Use of the reflective type of attenuator has been found to be important in enabling the gain in either the in-phase or quadrature path to be any value in the range -1 to 0 to $+1$.

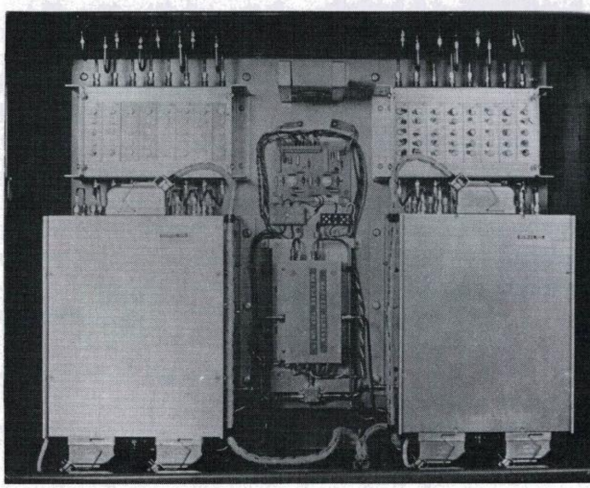


Fig. 12. The masthead box. The r.f.-input connections are at the top of the photograph with the sixteen selective filters mounted as two blocks of eight near the top. The two large enclosures contain the blocks of eight complex attenuators.

automatic locking circuitry and a capture range $>$ 10 times the maximum input frequency error which can occur.

The main features of the modified receiver are:

(i) A modified synchronous detector with locking circuitry sufficiently powerful that if line synchronising pulses are provided, the detector will lock in the presence of noise of CCI of level equal to the wanted signal.

(ii) A synchronising pulse separator which operates

off an envelope detector and which is capable of correct separation of stable line synchronising pulses for noise and CCI levels up to ~5 dB lower than the wanted signal.

(iii) Special measuring circuits.

This combination ensures a receiver capable of locking to the correct signal, and thus providing correct outputs from the measuring circuits. Without these three features, the aerial would not be able to optimise and hence reject the interference. The block diagram of the receiver is shown in Fig. 13.

The co-channel interference is measured by sampling the synchronising pulse tips of the video, a signal which, in the absence of interference, should be constant in level. Interference appears as a beat at the synchronising pulse tips and is extracted by sampling to give a complex low-frequency signal in the frequency ranges 0–1 kHz or 2–4 kHz; these frequencies resulting from the aliasing of the possible offsets of the transmission frequencies used in the United Kingdom to reduce visual patterning caused by co-channel interference. These sample signals are fed into the measuring system.

(d) Measuring system

Detailed theoretical and experimental work has shown that one of the major problems of adaptive systems, and the adaptive aerial in particular, lies in the area of measurement. A simple measuring system might operate by measuring the amplitude of the CCI beat, changing the aerial pattern slightly and remeasuring all amplitude of the beat. This was tried and worked well for CCI caused by unmodulated signals. It does not work satisfactorily, however, for real interference whose modulation is changing, even when the measuring period is synchronised to the frame period. For this reason, an alternative system was designed. This operates by defining two aerial patterns *A* and *B* with a very small difference in their control settings called the perturbation. The aerial is switched between these patterns on a line by line basis. The *A* alternate synchronous samples now provide a measure of the CCI for the *A* pattern and the *B* synchronous samples for the *B* pattern. Unfortunately, however, in the system described here, the output is in continuous use for rebroadcast and therefore any changes provided to the aerial must not

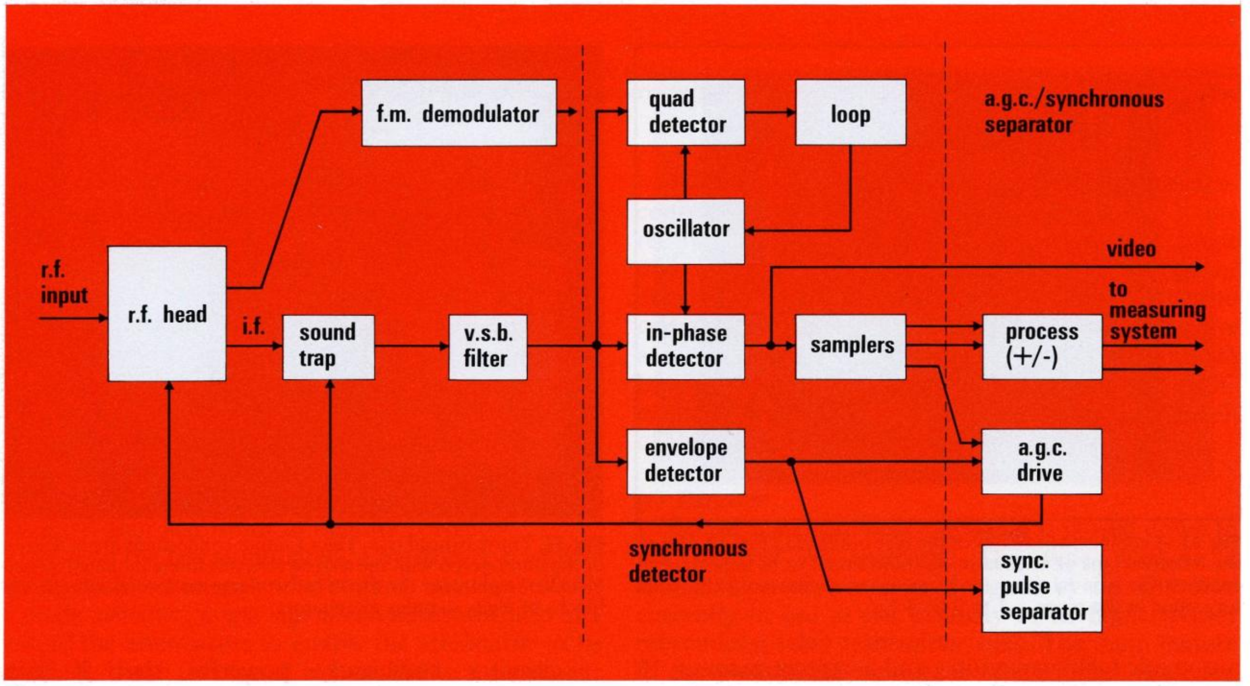


Fig. 13. Block diagram of the receiver. This is based on a standard IBA receiver, incorporating synchronous demodulation to eliminate the quadrature distortion that results from envelope demodulation of a vestigial-sideband filtered signal. Modifications to the standard receiver include more powerful locking circuitry; a synchronising pulse separator operating from an envelope detector capable of providing stable synchronising pulses with noise or CCI levels only 5 dB down on the wanted signal; and the special measuring circuits for adaptive control shown in Fig. 14.

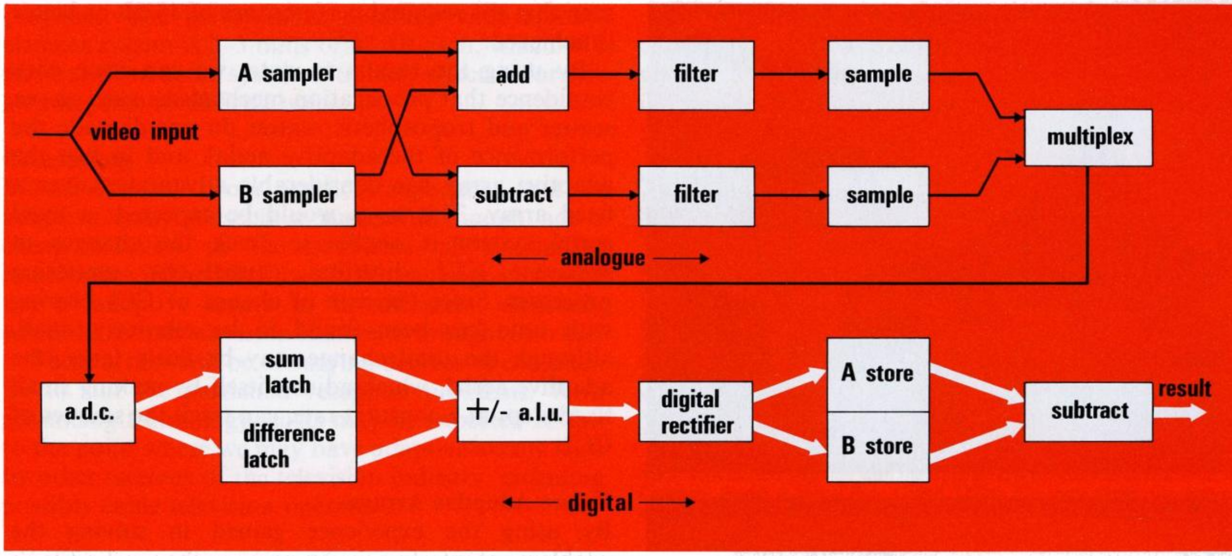


Fig. 14. The CCI measuring system used in the receiver to provide adaptive control of the aerial array. The processed analogue signals are converted to digital form for further logic processing to provide the control signals.

be visible on the output. The perturbation size is set such that for CCI levels up to equal to the wanted signal level incoming to the receiving site, the CCI can be rejected to greater than ~ 56 dB down on the wanted signal on the final output (carrier-to-carrier ratio). This very small perturbation necessitates a very accurate measuring system capable of detecting changes greater than 0.2% in the CCI level, especially for multiple CCI. The method used is shown in Fig. 14.

The analogue filters provide a carefully tailored response to eliminate for example aircraft flutter, and also to provide a suitable weighting of the nonoffset 0–1 kHz interference with respect to the offset interference (aliased to 2–4 kHz), since nonoffset interference is subjectively far more annoying for a given level than offset. The major feature of the above arrangement which is in essence filtering and detecting the *A* and *B* signals is the use of summing and differencing prior to the filters. This eases the filter component tolerancing from very difficult to simple and, in fact, abnormal differences in the filters result merely in desensitising the detector. This circuit has worked well in operation and indeed has a sensitivity $\sim 0.1\%$, the limits being set by the initial samplers and the digitising accuracy. Digital detection and integration is used because it is very difficult to produce analogue detectors with the required accuracy and reproducibility.

(*e*) *Control unit*

Adaptive control is provided by hardwired logic. Many algorithms were investigated during our computer simulations and it was found that the simple 'hill climb' modified to maintain constant gain, combined simplicity with a performance only about half the speed of very sophisticated and complex algorithms requiring large amounts of computation. Preset control values are incorporated so that during the night, for example, the aerial pattern is fixed ready for optimisation when the wanted signal is turned on. The controls are fed to the masthead box via analogue drivers which incorporate a degree of lightning protection.

Operational Performances

The first operational system has now been in service for several years on Alderney. It has helped the IBA and more recently the BBC to maintain a virtually continuous colour service without dropout caused by excessive CCI. Experimentation and measurements indicate that the aerial system is capable of rejecting up to twelve or so independent sources of interference. Note that independent channel frequencies of the same transmitter location correspond to independent sources. Figure 15a shows the effect of a picture of co-channel interference as obtained when the SABRE aerial pattern is held at its preset value and Fig. 15b shows the result after

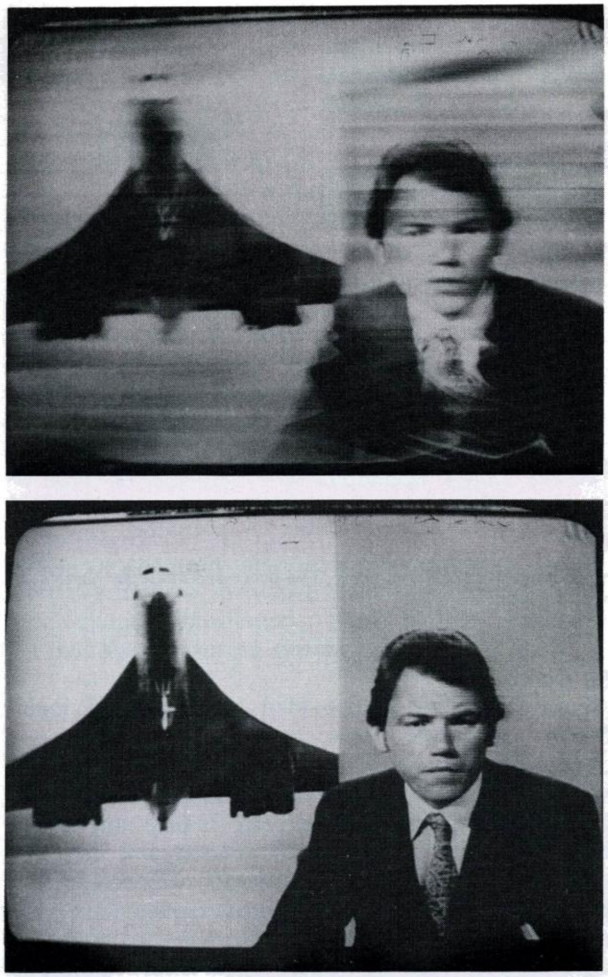


Fig. 15. Unretouched off-screen photographs of pictures obtained at Alderney using the SABRE array under conditions of high level co-channel interference: (*a*) before optimisation: (*b*) after optimisation. To obtain (*a*) the SABRE aerial pattern was held at its preset value. The CCI rejection improves from about 20 dB in the preset pattern to around 45–50 dB in the optimised pattern.

optimisation of the aerial. The improvement in CCI rejection is from around 20 dB for the preset pattern to around 45–50 dB for the optimised pattern.

Measurements have also been made during 1978–79 of the levels of interference on the output of the adaptive array and also of a fixed aerial array used as a reference. Results for a period of relatively high interference are shown in Fig. 16.

It can be seen that, under these conditions, the adaptive array achieves a performance 25 dB better than the fixed array and, if the rejection of the latter to interference is assumed to be 20 dB, then it can be seen that the required performance of 45 dB or better is achieved.

By using the results to date, we can state with confidence that propagation mechanisms such as sea scatter and tropospheric scatter do not degrade the performance of the adaptive aerial, and in fact the adaptive array has considerable advantages over a fixed array. This is as would be expected; a fixed aerial system is unable to track the changes in apparent CCI direction caused by scattering processes. Since the rate of change of CCI bearing with time has been found to be relatively small, although the total change may be fairly large, the adaptive aerial, which adjusts itself by making small fixed steps at a constant rate, can track the sources of CCI.

Future Adaptive Arrays

By using the experience gained in solving the problems of interference on perhaps the most difficult UHF link in the United Kingdom, the IBA is investigating the design of a simpler four-element system for use when there are fewer sources of interference. The principles will remain the same, but with four elements the amount of equipment is considerably reduced, and this, combined with rapidly improving technology, makes possible the use of a microprocessor to control the array.

A microprocessor-controlled array will have the power not only to operate as an RBR, but also as a monitoring array in which the pattern can be switched to optimise on different transmitters for monitoring

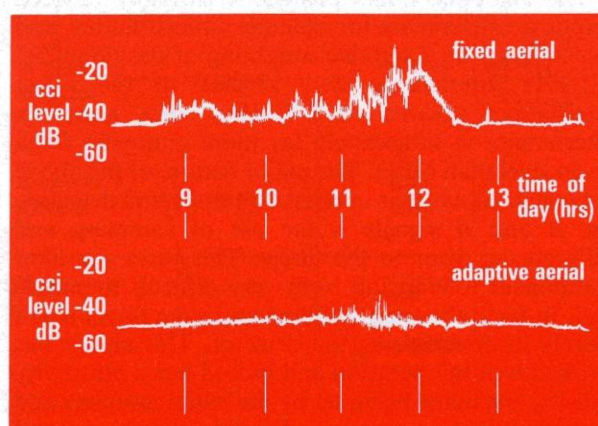


Fig. 16. Plots of co-channel interference levels relative to the wanted signal as a function of time. *Above* plot using a high-gain, high-directivity fixed array. *Below* plot for the SABRE array. The measurements relate to a period of relatively high interference.

purposes. The performance obtained with a four-element system is 2–3 nulls of 45 dB each, which will make such an aerial system extremely effective for off-air reception in difficult areas.

Conclusions

We have shown from practical experience that an adaptive array presents a successful though not a cheap solution to the problems of co-channel interference in the UHF band. We envisage increased use of adaptive arrays by broadcasters, especially when the fourth television channel is brought into use.

These arrays will be of varying degrees of complexity to suit the particular reception problems. With advancing technology, costs may be expected to fall to the point where we may have adaptive arrays used by other sections of the television industry, including possibly cable television operators.

Finally, the adaptive array represents an important concept in which the aerial and the receiver are not treated as separate items to be considered independently, but as intimately related parts of a whole receiving system.

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Note: This section, together with the Appendix, is basically as in *IEE Proc. F, Commun., Radar & Signal Proc.* (1980), **127(4).** For these papers the authors were awarded the IEE's 1981 'Blumlein-Browne-Willans Premium'.

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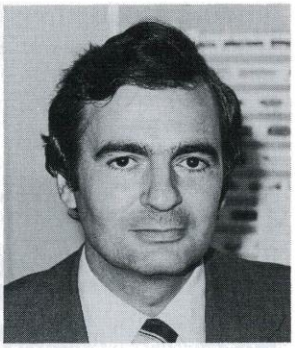


CCI Suppression Techniques

by J. S. Lothian

Synopsis

This paper describes some results of a detailed investigation into signal-processing techniques for discriminating against 'offset' co-channel interference (CCI) by means of adaptive comb filtering. The basis for comb filtering of a television signal lies in the fact that the luminance energy is concentrated around harmonics of the line scanning frequency (15.625 kHz for System I). Offset operation of transmitters causes the two spectra to become interleaved but can be separated by means of a comb filter, that is to say a transversal filter with line-period (64 μsec) delay lines. The main features of an experimental filter are described and its performance indicated under different conditions and modes of operation. The comb filter interfaces with a receiver using synchronous demodulation and permits the simultaneous cancellation of interfering carriers on both offsets, together with the low-frequency luminance components of the interfering signals. It is shown that comb filtering is technically feasible and capable of providing an improvement of up to two 'picture grades' (roughly 12 dB improvement in co-channel protection ratio) provided that the unwanted carriers are 'offset' from the wanted carrier. A complex, adaptive comb filter employing in-phase and quadrature signals can reduce CCI on both 'offsets' and will also provide optimum protection from one source of interference. The filter can be interfaced with existing synchronous-demodulator receivers and involves no changes to the RBR aerial installation. The comb filter thus provides a further anti-CCI weapon in the armoury of the systems design engineer, complementary to the more elaborate adaptive aerial systems that may be necessary where severe CCI is likely or where the interference is from transmitters that are not 'offset' in carrier frequency.

UHF television service area planning engineers have found when investigating sites for off-air RBR links to provide programme feeds for relay stations that the most common problem is that presented by interference from co-channel transmissions. The strength and duration of such interference may be of varying duration and magnitude depending on distance, intervening terrain, radiated power and the directivity and polarisation discrimination of the receiving aerial. Some estimate may also need to be made of the likely incidence of anomalous propagation, particularly where sea paths are involved.

Much of the work of minimising CCI falls within the responsibilities of the planning engineers who have accumulated much experience in this field (*IBA Technical Review Vol. 7*). The full effect of co-channel interference could be measured at each prospective relay station by long term field strength recordings, extending over a period of a year or more. In practice, estimates are made using a computer and taking into consideration existing and planned high-power transmitters in the UK and the Continent as well as lower-power relay stations in the UK. Relay coverage and RBR links may then be planned on the basis of 'protected field strength' for 95% or 99% of the time.

It is necessary to ensure that all re-broadcast receiver (RBR) sites have a protected field strength for the wanted signal exceeding the value of 80 dB/μV/m for a main station or 76 dB/μV/m for a lower-power relay. It is often necessary to check by measurement the signal level(s) that are received at the site from the principal potential CCI source(s).

To combat CCI the standard remedies involve careful selection of channel grouping for all stations and utilising of polarisation discrimination. Normally in the UK 'main' UHF stations use horizontal polarisation and lower-power relay stations use vertical polarisation, but this may be varied in some cases, particularly where it is desired to minimise tidal fading in areas served across tidal waters (vertically polarised signals suffer less tidal fading than horizontally polarised signals).

One or more unwanted signals can be minimised by increased directivity of the RBR receiving aerial: this is facilitated where the aerial has a regular radiation pattern (Yagi aerials may have relatively poor front-to-back ratio but have deep side lobes; log-periodics may have less forward gain but a smoother polar diagram with good front-to-back ratios). Two or more arrays may be connected together through phasing lines arranged to provide deep nulls in the direction of potential sources of CCI. However, as shown in an earlier section, there are practical limitations to the null depth of fixed arrays mounted on a high mast, owing to, for example, mechanical movement of aerials in conditions of high winds.

The adaptive aerial array forms an elegant, but expensive, solution in those situations where a number of potential sources of CCI exist without the use of frequency offsets. In this section we consider primarily the situation where frequency offsets exist or can be provided.

Comb Filtering

Where the spectra of wanted and unwanted signals can be distinguished it becomes possible to use comb filtering of the intermediate frequency or baseband frequency[1]. Since such a technique cannot reject interference from a non-offset transmitter (and introduces some waveform degradation) it cannot provide a solution for the Alderney-type situation.

The IBA work has concentrated on the use of a comb filter to discriminate against CCI on broadcast television signals when interfacing with an off-air receiver employing synchronous demodulation. In these circumstances:

(i) Comb filtering can give an improvement of 12 dB in the required protection ratio of off-set CCI.
(ii) This degree of improvement can be achieved in situations where a single source of interference predominates by the use of a fixed filter.
(iii) Where interference exists on both off-set spectra, an adaptive system is required for optimum performance.

Comb filtering thus has potential applications to RBR links and also to the visual monitoring of the technical quality of pictures received from distant stations. It is inevitable that as the number of new UHF transmitters increases, for example for Channel Four, the number of sites affected by CCI, even when normal planning precautions have been taken, will increase. It seems likely that in a number of cases the CCI problem could be eased by operational use of comb filtering: in particular the comb filter provides protection against unwanted signals arriving along or from close to the bearing of the wanted signal, a situation that imposes severe problems when using aerial array nulls.

The fundamental basis for comb filtering lies in the fact that the luminance energy of a television signal is concentrated around harmonics of the line scanning frequency (15.625 kHz). Offset operation of transmitters causes the two spectra to become interleaved. This is represented in Fig. 1 for the case of a positive single offset. Clearly it is possible to separate the spectra by means of a comb filter and a transversal filter employing 64 μs delay lines will provide the required periodicity in the frequency response. To simplify matters we may consider the frequency response between f_w and (f_w+f_L) and refer all interfering signals to this region. Thus a notch is required at $(f_w+\frac{1}{3}f_L)$ for offsets of $-\frac{5}{3}$ and $+\frac{10}{3}$, at $(f_w+\frac{2}{3}f_L)$ for offsets of $+\frac{5}{3}$ and $-\frac{10}{3}$. Without precision offset operation, frequency errors in transmitters and transposers will cause the offset frequency to deviate from the nominal—Fig. 2 shows a typical probability distribution. Thus a system designed to deal with interference on both offsets must provide two discrete notches which may be independently adjusted in frequency.

THE COMPLEX COMB FILTER. Using synchronous demodulation[2] it is possible to obtain simultaneous in-phase and quadrature video outputs. The complex comb filter, shown in Fig. 3, makes use of both of these signals in order to allow independent control of two notches to give optimum cancellation. Normally of course, the video output would be taken from the in-phase detector, the quadrature detector simply

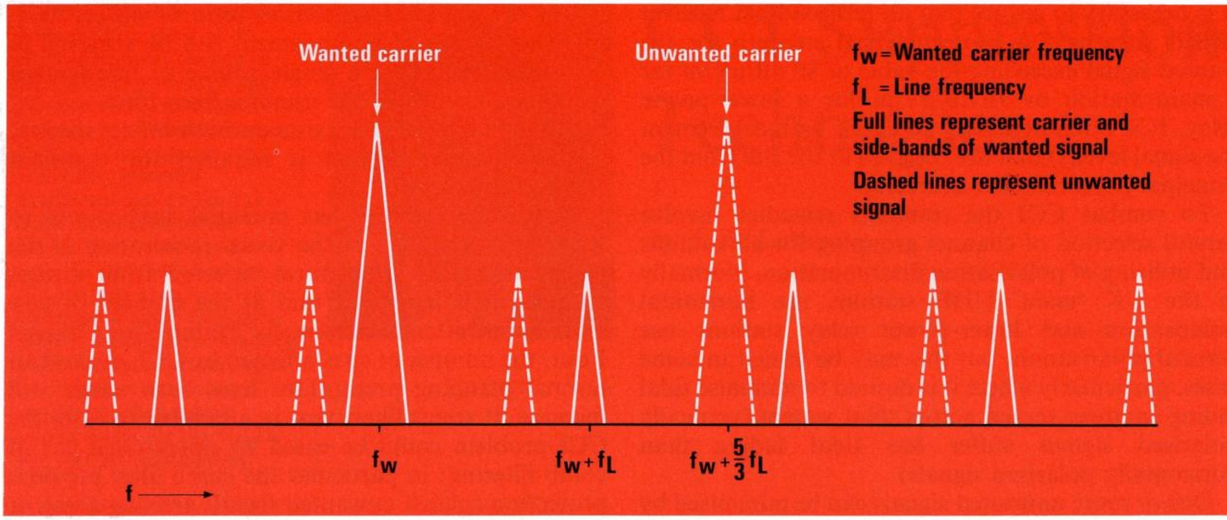


Fig. 1. Spectra of wanted and unwanted signals in the region of a wanted carrier (f_w) and an 'offset' unwanted carrier ($f_w + (5/3f_L)$). The effectiveness of comb filtering stems from the concentration of the luminance energy around harmonics of the line scanning frequency (f_L) which for 625-line 50-field systems is 15.625 kHz. Off-setting a carrier by a suitable fraction of f_L causes the two luminance spectra to be interleaved throughout the video bandwidth.

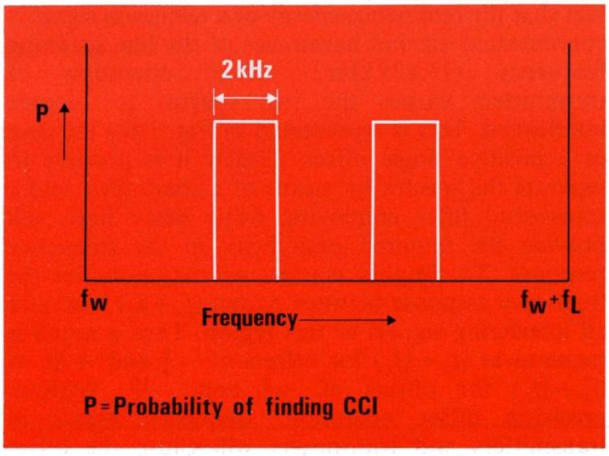


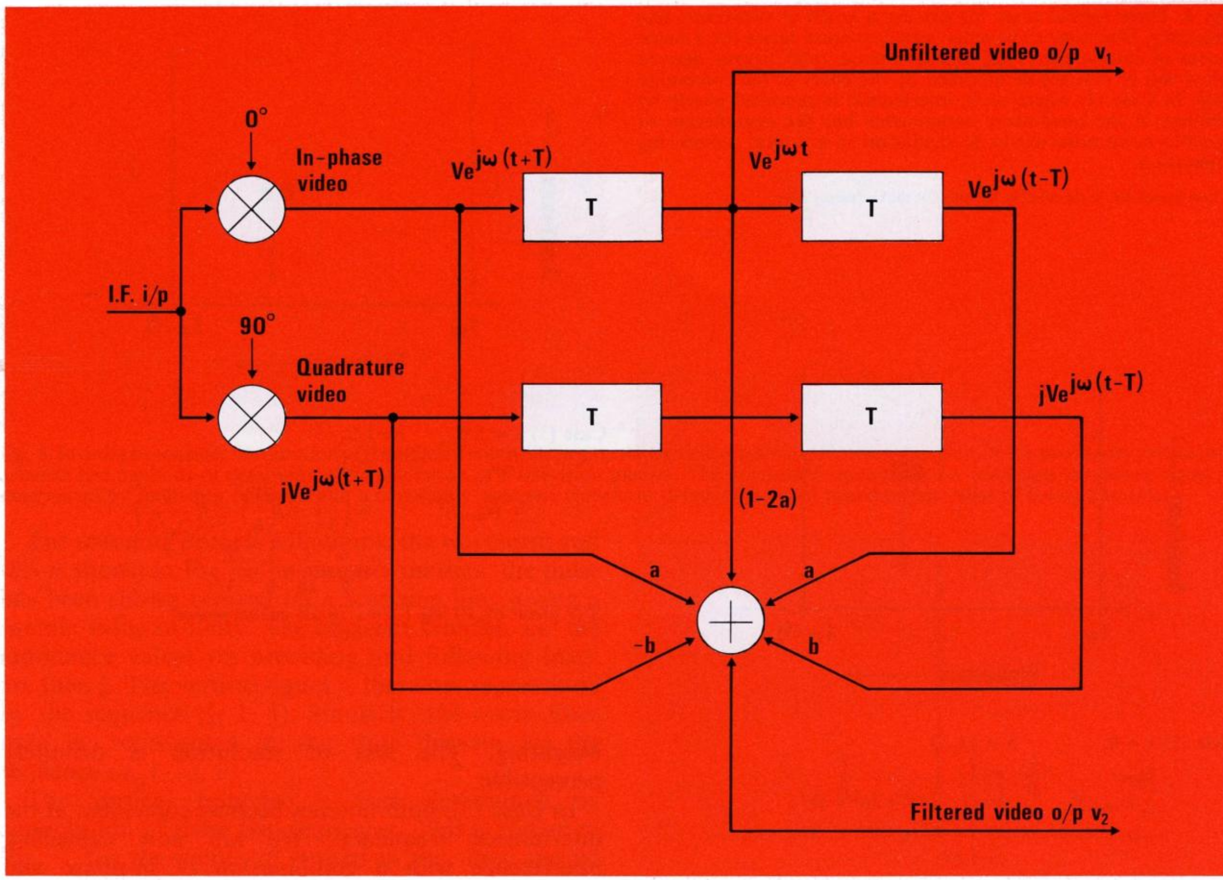
Fig. 2. To maintain accurate interleaving of the spectra from an unwanted transmission requires that both carriers have a high order of frequency stability. For 'precision offset' carriers are maintained within a tolerance of ± 2.5 Hz but normal practice in the UK permits a tolerance of ± 1 kHz, and offsets at $-(5/3)f_L + (10/3)f_L$ require a notch at $f_w + (1/3)f_L$ while offsets of $+(5/3)f_L$ and $-(10/3)f_L$ require a notch at $f_w + (2/3)f_L$. The permitted frequency tolerances in transmitters and transposers result in a probability distribution of finding CCI as shown.

acting as a phase-sensitive detector for phase-locking the reference oscillator.

The choice of the filter coefficients a and b not only determines the two notch frequencies, but has a significant effect on the wanted video signal. The in-phase signal at the outer tapping points contains the wanted video signal, delayed or advanced by one line. Thus we have a loss of vertical resolution of the picture, due to averaging the wanted signal over three lines, and this loss increases as we raise the value of a.

The quadrature signal contains only high frequency components of the wanted signal. This is a consequence of the vestigial sideband modulation system used for television transmission. Low frequency components of the wanted signal are suppressed, at the quadrature detector output, due to the fact that both sidebands are present in anti-phase. Further cancellation takes place because the outer tap signals in the quadrature path are subtracted. Hence any quadrature component of the wanted signal that is line repetitive will be cancelled. We can therefore increase the coefficient b without causing loss of vertical resolution. This benefit is offset by a degradation in video signal-to-noise ratio, since the quadrature component suppresses signal but not noise.

In Fig. 4 are shown three special cases for placing a notch at two-thirds line frequency. Case (1) has $a = 0$, the interference being cancelled purely by addition of quadrature components. The theoretical reduction in signal-to-noise is 2.2 dB and, in practice, subjective measurements have given figures of 2 to 3 dB. This degradation is significant in a typical monitoring application, where the incoming signal-to-noise ratio may already be marginal. Conversely, Case (2) with $b = 0$ actually improves the signal-to-noise as a result



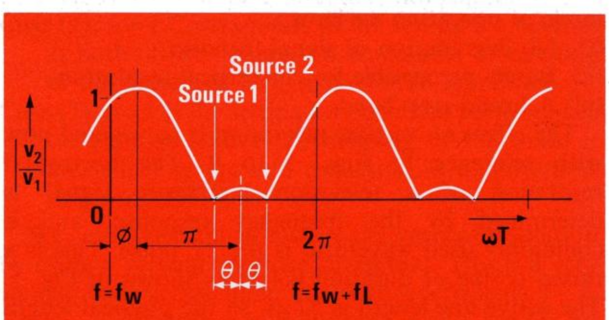


Fig. 3. (*Above*) Complex comb filter making use of simultaneous in-phase and in-quadrature video outputs obtained from synchronous demodulation of a signal at intermediate frequency. This permits independent control of two notches determined by the filter coefficients a and b. (*Left*) Determination of the null frequencies:

$$v_1 = V\,e^{j\omega t}$$

$$v_2 = (1-2a)V\,e^{j\omega t} + a[V\,e^{j\omega(t+T)} + V\,e^{j\omega(t-T)}] - jb[V\,e^{j\omega(t+T)} - V\,e^{j\omega(t-T)}]$$

$$\frac{v_2}{v_1} = 1-2a+a(e^{j\omega T}+e^{-j\omega T}) + \frac{b}{j}(e^{j\omega T}-e^{-j\omega T})$$

$$= 1-2a+2a\cos\omega T+2b\sin\omega T$$

$$= 1-2a+2k\cos(\omega T-\phi) \text{ where } \phi = \tan^{-1}(b/a)$$

$$k = \sqrt{a^2+b^2}$$

Condition for a null is

$$\cos(\omega T-\phi) = -\left(\frac{1-2a}{2k}\right) = \cos(\pi\pm\theta)$$

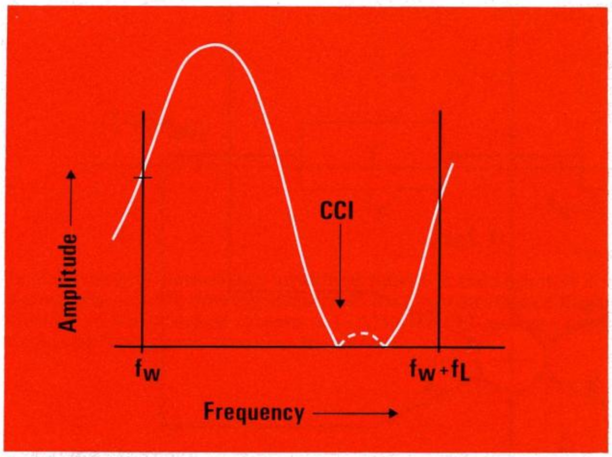
$$\text{where } \cos\theta = \left(\frac{1-2a}{2k}\right)$$

$\therefore$ Nulls occur at $\omega T = \phi+\pi\pm\theta$

of signal averaging. This improvement figure would only be true for a line repetitive signal—in general it is somewhat less. This mode of operation allows cancellation of both offsets, assuming no errors in the transmitter frequencies. Note however, that we are adding three consecutive lines together with equal

Fig. 4. Three special cases for placing a notch at two-thirds line frequency. Diagrams show video signal-to-noise ratios for a single source of interference. The signal-to-noise ratio figures derived below will in practice be modified by the effect of the combining filter. In Case (1) where $a=0$, interference is cancelled purely by addition of the quadrature components but the degradation of signal-to-noise ratio would be significant in a typical monitoring application.

* These figures will be modified by the effect of the combining filter.

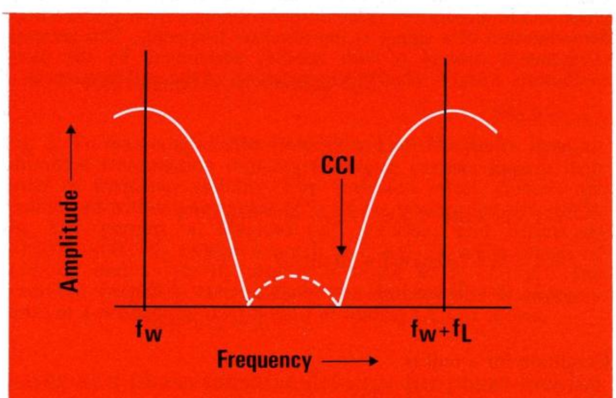


Case (1) $a = 0$, $b = 1/\sqrt{3}$

$$v_{\text{signal}} = 1$$

$$\sum (v_{\text{noise}})^2 = (1-2a)^2 + 2a^2 + 2b^2 = 1 + 2(1/\sqrt{3})^2 = 1.67$$

S/N given by 10 log 1.67 = 2.23 dB *degradation**

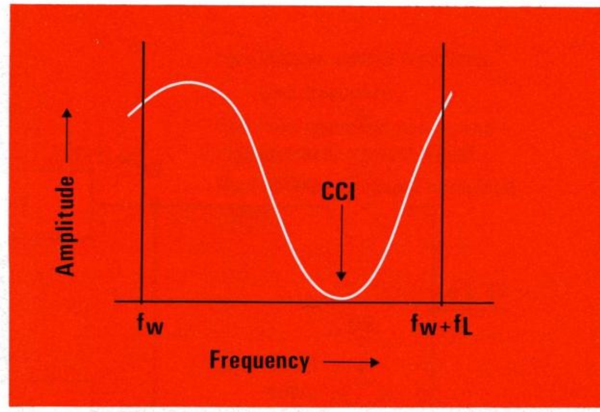


Case (2) $a=\frac{1}{3}$, $b = 0$

$$v_{\text{signal}} = 1$$

$$\sum (v_{\text{noise}})^2 = 3(\tfrac{1}{3})^2 = \tfrac{1}{3}$$

S/N given by 10 log 3 = 4.77 dB *enhancement**



Case (3) $a = 1/6$, $b = 1/2\sqrt{3}$

$$v_{\text{signal}} = 1$$

$$\sum (v_{\text{noise}})^2 = \left(\frac{2}{3}\right)^2 + 2\left(\frac{1}{6}\right)^2 + 2\left(\frac{1}{2\sqrt{3}}\right)^2 = \frac{4}{9} + \frac{1}{18} + \frac{1}{6} = \frac{2}{3}$$

S/N given by 10 log 1.5 = 1.76 dB *enhancement**

weighting. The loss of resolution is definitely perceptible.

In Case (3), both notches become coincident at the interference frequency. We are now cancelling interference with a combination of in-phase and quadrature signals and this setting has been found in practice to be optimum for a single source. Under this mode of operation we have:

(i) No degradation of signal-to-noise.
(ii) Barely perceptible loss in vertical resolution.
(iii) A broad stop band.

The effect on vertical resolution is considered now with reference to Figs. 5 to 9. The horizontal resolution of a television luminance signal is determined by the frequency response, and a commonly used waveform is the $2T$ pulse. This is a raised cosine of duration 0.4 µs as shown in Fig. 5. The active line length is 52 µs, so the physical length of the pulse is $0.4\,h/52$ where h is the horizontal picture dimension. Now consider a waveform in the vertical direction having the same spatial properties. The number of picture lines per field is given by $625/2-25=287.5$. Hence, we can express the length of the waveform in the vertical direction in terms of line periods:

$$0.4\,h/52 \times 4/3\,h \times 287.5 = 2.95 \simeq 3$$

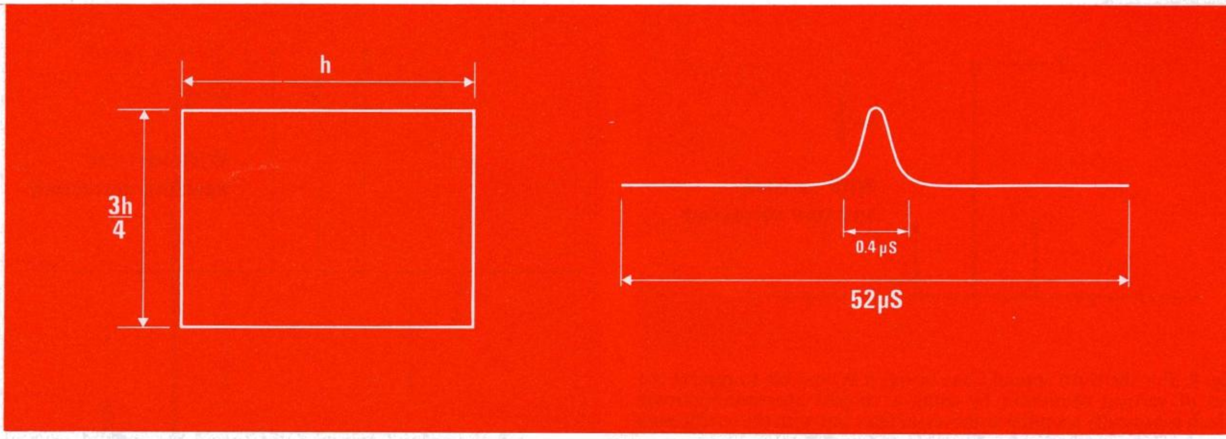


Fig. 5. In order to consider the question of the effect of comb filtering on the vertical resolution of a television picture, it is necesssary to take into account test methods of determining the resolution of a television picture. The horizontal resolution of a television luminance signal is determined by frequency response and a commonly used test waveform is the $2T$ pulse, a raised cosine pulse of 0.4 µs duration.

The scanning process will sample the waveform and this is shown in Fig. 6. To simplify matters, the pulse has been shown centred on a scanning line, giving a sample value of unity. The adjacent samples, i.e. the luminance values on preceding and following lines, are then $\frac{1}{4}$. The vertical signal is therefore represented by the sequence ($\frac{1}{4}$, 1, $\frac{1}{4}$). Similarly, the comb filter may be represented in the time domain by the sequence (a, 1–2a, a).

The vertical response is then determined by convolution of these two sequences. This is shown in Fig. 7, for the values of a used in the three special cases of Fig. 4.

Case (1): $a=$o We have of course no degradation.
Case (2): $a=\frac{1}{3}$ Corresponds to the setting for notches at both offsets. There is considerable broadening of the pulse and 50% reduction in amplitude.
Case (3): $a=\frac{1}{6}$ Is the setting for optimum performance on a single source. Broadening of the pulse is much less severe and reduction in amplitude is now only 25%.

We could attempt to restore the vertical resolution by using a vertical aperture corrector[3]. This is a very similar transversal filter but with coefficients ($-x$, $1+2x$, $-x$) where x is adjustable. For Case (3) we require $x=0.3$ and the overall response is as shown in Fig. 8. This is very close to the original waveform showing that we could, if required, restore the signal in Case (3). Similar calculations for Case (2) show that, although we can restore the amplitude by means of very large amounts of vertical aperture correlation, the pulse broadening is not reduced and large overshoots are produced.

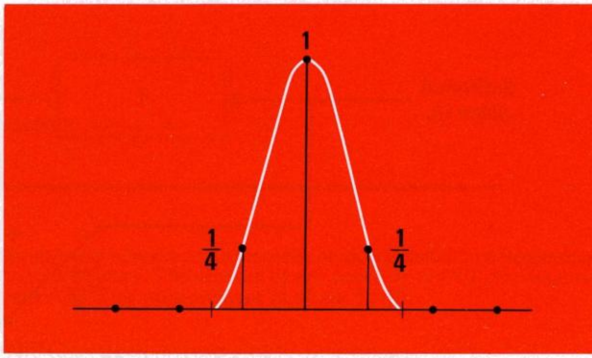


Fig. 6. A $2T$ waveform expressed in the vertical direction in terms of line periods sampled by the scanning process and simplified by centring the pulse on a scanning line to give a sample value of unity. Adjacent samples (luminance values) on preceding and following lines are one-quarter and the vertical signal is represented by the sequence $\frac{1}{4}$, 1, $\frac{1}{4}$ and the comb filter may be represented in the time domain by the sequence a, 1-2a. a.

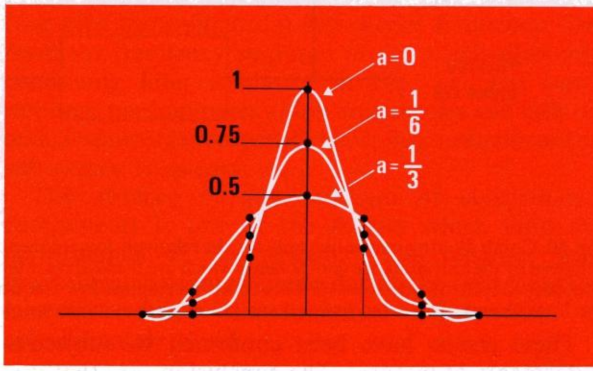


Fig. 7. Vertical response determined by convolution of the $2T$ pulse and the comb filter for the three special Cases where a is 0, $\frac{1}{3}$rd and $\frac{1}{6}$th.

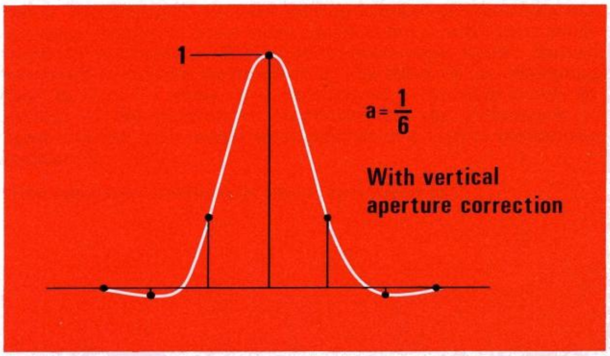


Fig. 8. For the third special Case ($a=\frac{1}{6}$) it is possible to restore the loss of vertical resolution by using a vertical aperture corrector which is basically a similar transversal filter but with the coefficients $-x$, $1+2x$, $-x$ where x is adjustable and is about 0.3. Thus for the third Case the relatively small loss of vertical resolution is subjectively not disturbing but could be corrected if required.

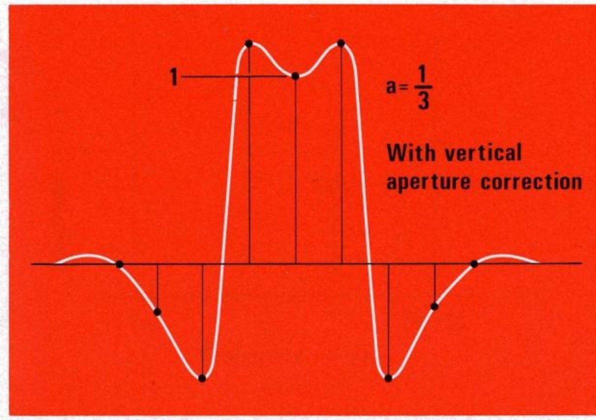


Fig. 9. For the second special Case ($a=\frac{1}{3}$) the loss of vertical resolution is noticeable and cannot be corrected by vertical aperture correction. Such loss of vertical resolution would not be acceptable for visual monitoring of the quality of transmissions but would be acceptable for most other purposes.

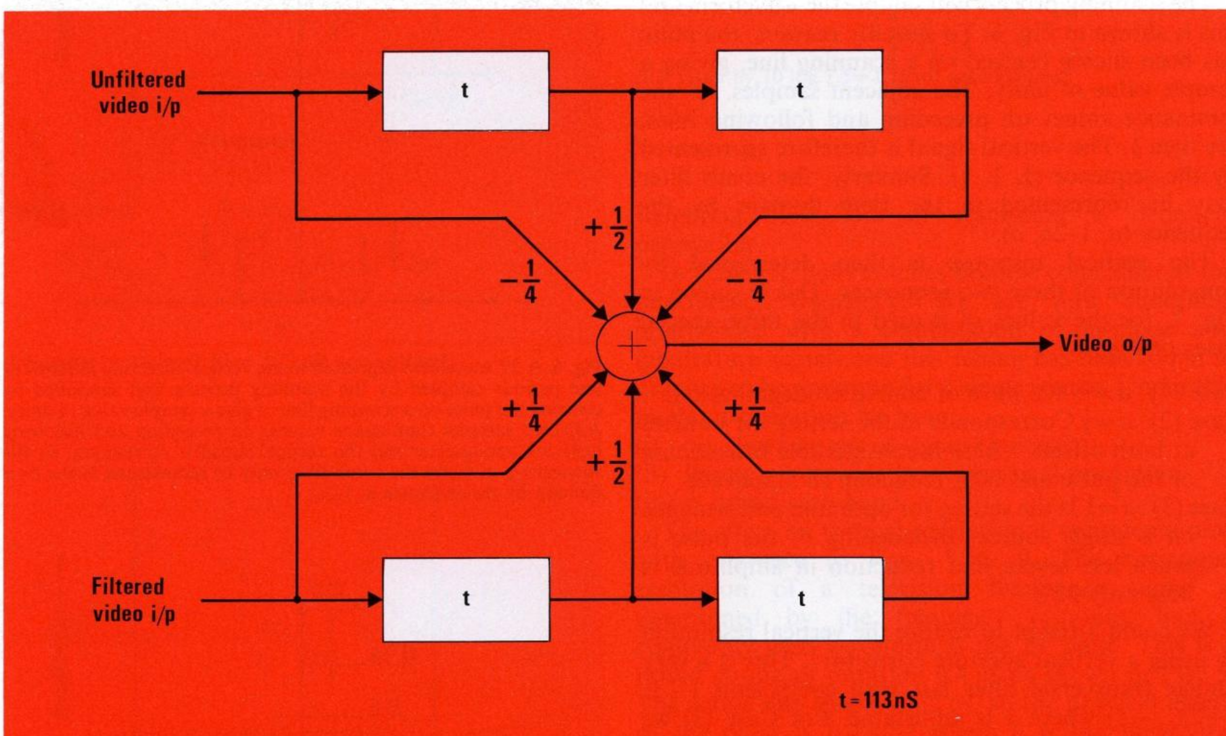


Fig. 10. Comb filtering is effective only for the relatively low frequency luminance components of the video signal and cancellation cannot be achieved of the chrominance components which are themselves interleaved with the higher frequency luminance components. It is therefore necessary to split the video band prior to comb filtering and low-frequency components and then to provide a complementary combining filter of the form shown. The delay lines (t) are 113 ns with suitable weighting coefficients.

These results have been confirmed by subjective judgements of pictures. The conclusions are that it is important, where possible, to restrict the value of a in order not to degrade vertical resolution. For $a=\frac{1}{6}$ the loss is small, subjectively not disturbing and could be corrected if required. For $a=\frac{1}{3}$ the loss is noticeable and cannot be corrected.

From these considerations it appears that the comb

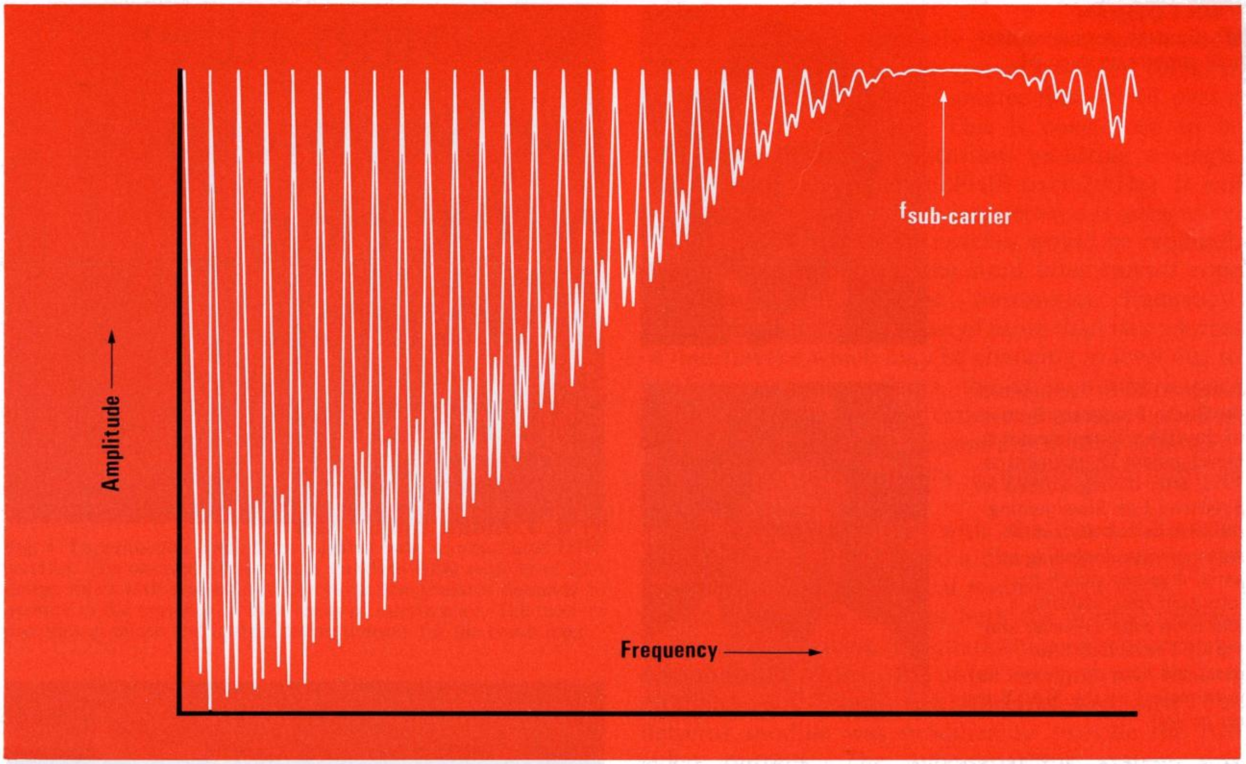


Fig. 11. Resultant frequency response of the comb filter and combining filter showing how the comb response predominates at low frequencies but falls off with a sinusoidal response, becoming insignificant at sub-carrier frequency. The absence of any rejection of CCI chrominance components is one of the limiting factors of the system, restricting overall subjective improvement to about 12 dB.

filter will deal with one interfering source to re-broadcast standards of quality, and with two sources on different offsets to a slightly lower standard suitable for visual monitoring at Regional Operations Centres.

COMBINING FILTER. In describing the signal processing which takes place in the comb filter, it has been assumed that we are dealing with low frequency luminance components of the video signal. Chrominance components in the PAL system occur at $(n \pm \frac{1}{4})\, f_L$ and will not be cancelled by this system. Furthermore, the wanted chrominance will be degraded by the comb filter, and it is therefore necessary to split the video band into two parts, combining low frequency comb-filtered components with unfiltered high frequency components. Complementary filters are required having the following properties:

(i) Accurately matched amplitude/frequency response.
(ii) Equal time delay.
(iii) Constant group delay response.
(iv) Symmetrical response with respect to sub-carrier frequency.

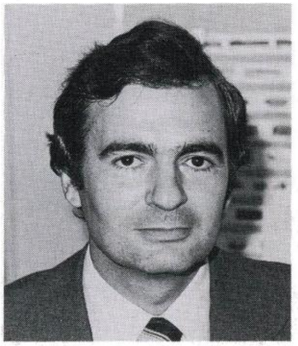
This has been achieved with transversal filters of the form shown in Fig. 10. The delay lines in this case are 113 nsec with weighting coefficients chosen to give in one case zero transmission at dc, and in the other case zero transmission at sub-carrier frequency. The resultant frequency response of the comb filter and combining filter is shown in Fig. 11. The comb response predominates at low frequencies and falls off with a sinusoidal response, becoming insignificant at sub-carrier frequency.

The absence of any rejection of chrominance components is one of the factors which limit the performance of the system. This, together with breakthrough of interference during the field blanking interval, restricts the subjective improvement to approximately 12 dB.

References

Selected references relevant to both this Section and the next will be found at the end of the next Section.

JOHN LOTHIAN, MA(Cantab). A biographical note appears on page 24



A Comb-filter for Suppression of CCI on Television Signals

by J. Lothian and J. Airs

JOHN AIRS B. Tech. joined the Radio Frequency Section of the IBA's Experimental & Development Department in 1975, after having worked for Solartron Ltd. Since joining the IBA he has worked on many projects including an aerial diversity switch for television relay stations, a television echo canceller and the CCI reduction filter described here. At present he is employed on the MATE test set project.

Synopsis

This paper describes the practical performance of a prototype comb filter designed and built along the lines suggested in the preceding Section. The experimental filter was subjected to laboratory testing using simulated CCI signals generated by two modified IF test transmitters and also to a short field trial at an IBA mobile maintenance base where subjective observations were made under field conditions of CCI.

These trials confirmed the value of such filtering as a means of improving the CCI protection ratio though imposing a penalty of some loss of vertical resolution.

These tests also suggested that for field use it would be desirable to simplify and reduce the number of controls, and a simplified filter has been developed. The authors also believe it may be possible to implement the filter using alternative delay elements, for example a shift register, RAM store or charge-coupled device.

A prototype comb CCI-suppression filter capable of providing 12 dB suppression to single sources of offset CCI was designed and built by IBA engineers based on the principles outlined in the preceding section. This filter has been subjected to laboratory and field testing, achieving the performance objectives but at the same time indicating a number of ways in which an operational filter of this type could be improved still further.

The comb filter, together with its associated UHF receiver, is shown in Fig. 1. The receiver is based on a standard re-broadcast receiver of IBA design and has been fitted with a synchronous detector designed to operate in the presence of high levels of interference. In-phase and quadrature outputs are fed to the comb filter unit.

The 64 μs delay lines are contained in two video delay modules, one of which is shown in Fig. 2. Each consists of two ultrasonic glass delay lines together with associated modulation and detection circuitry. Automatic gain control is used to compensate for change in tranmission loss with temperature. (Change

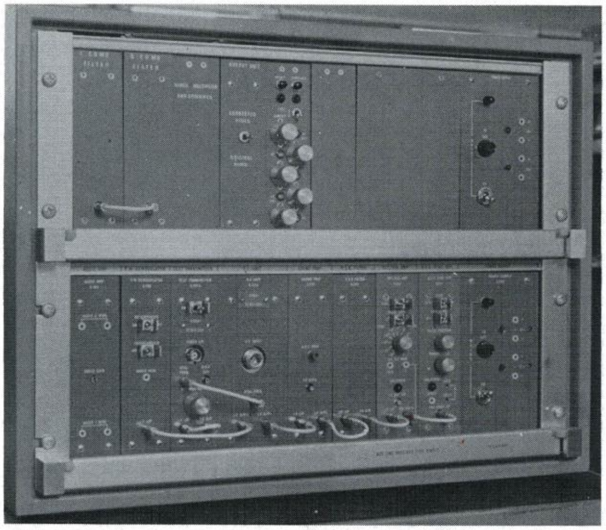


Fig. 1. Experimental comb filter together with its associated UHF receiver. The receiver is based on a standard RBR receiver of IBA design fitted with a special synchronous demodulator designed to operate in the presence of high levels of interference. The receiver provides in-phase and in-quadrature outputs for the comb filter.

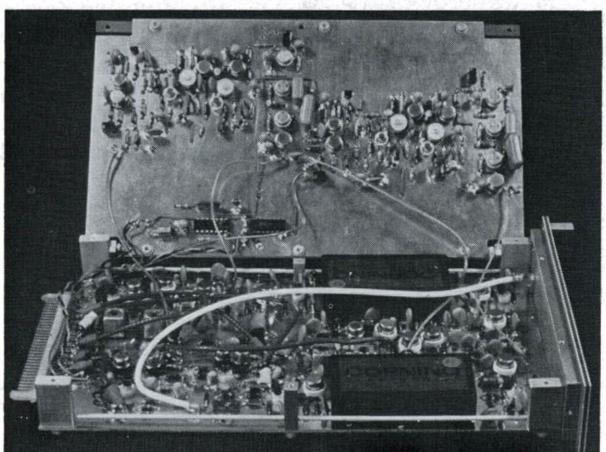


Fig. 2. One of the two video delay modules each of which incorporates two 64 μs ultrasonic glass delay lines. The unit also includes the associated modulation and detection circuitry with an AGC system to compensate for change in transmission loss through the delay lines with change of temperature.

in delay with temperature is minimised by the use of zero-temperature coefficient glass.)

The summing circuitry contains broadband four-quadrant multiplier circuits which allow the filter coefficients to be set up as analogue voltages, either manually or automatically, as part of an adaptive system.

As a fixed filter, manually preset to cancel a single source of interference, the system has been found to maintain a notch depth of greater than 30 dB over a wide temperature range. This is considered to be adequate for many applications requiring a single, broad notch (greater than 20 dB over 2 kHz). It will be appreciated that where interference is present on both offsets, two narrow notches would be required. Under these conditions the fixed filter cannot cope with any drift in the transmitter frequency; furthermore, in the presence of more than two sources of interference which may be changing in level due to UHF propagation effects, the filter would require continuous adjustment in order to maintain optimum performance. For these reasons any system capable of dealing with multi-sources of interference must be adaptive.

The results achieved with the fixed filter are described below, leading to a discussion covering the requirements for a practical multi-source filter.

The comb filter requires both the normal inphase video plus a source of quadrature synchronously-demodulated video. In order to provide this, a modified IBA Model E183 receiver was used. The detector module was modified to provide the two video outputs. The sync-separator system was designed to operate under high levels of CCI; this was originally developed as part of the SABRE project.

Laboratory testing of the filter was achieved using a CCI generator consisting of two modified IF test transmitters. The output of one transmitter, which was set to a carrier frequency five-thirds of line-frequency below the other, was fed, via an attenuator, to a summing junction where both transmitter outputs were added. Thus a source of single offset CCI at IF could be provided. Both transmitters could be modulated with a video signal via an internal black level clamp. The output of this generator was fed to the IF input of the E183 receiver.

The filter has five variable taps which can alter its frequency response and two different modes could be selected to null one source of single offset CCI. Both of these modes and their corresponding responses are shown in Fig. 4 of the preceding section. The first mode (marked 2) was the so-called 'narrow null' mode produced purely from inphase video signals ($a=\frac{1}{3}$, $b=0$); the second mode (marked 3) produced a single broad null and was designated the 'broad null' mode: ($a=\frac{1}{6}$, $b=\frac{1}{2}\sqrt{3}$). The two responses can be compared in Fig. 3. The tap values given are nominal and adjustments were made for the actual situation.

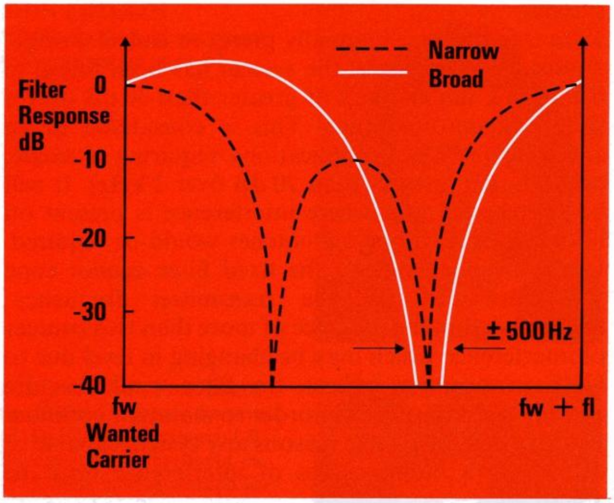


Fig. 3. 'Broad null' and 'narrow null' responses for simple filter.

Measurements

Two series of measurements were performed:

(i) Objective measurements in the laboratory of the filter operation with and without CCI present.

(ii) A short field trial at St. Hilary Mobile Maintenance Base where some subjective observations were made on the equipment's action with CCI present.

The first series of tests involved making photographic records of the picture improvements, due to the comb filter, as seen on a picture monitor. In order to observe subtle picture impairments, the Philips electronic test card was chosen to provide the wanted signal for the CCI generator. Interference at a relative level of −25 dB was produced by the CCI generator which was modulated by an off-air video programme signal, thus providing a realistic CCI signal. The CCI frequency was set to be exactly a single offset. A record was made of the degradations produced by the comb filter alone by turning off the CCI source. After changing the wanted signal to a line-repetitive pulse-and-bar waveform and removing the modualtion from the CCI source, the comb-filter output was recorded. The reason for using a line-repetitive waveform was so that the system impairments could be more easily observed. These tests were repeated for both the narrow- and broad-null modes of the comb filter. The laboratory tests simulated the expected CCI at the St. Hilary monitoring site when the Stockland Hill transmitter was being received.

A short field trial at St. Hilary was conducted in order to observe a typical off-air CCI problem, and to

TABLE 1: CO-CHANNEL INTERFERENCE OBSERVED AT ST. HILARY MOBILE MAINTENANCE BASE WHILE MONITORING STOCKLAND HILL, SOUTH DEVON ON CHANNEL 23+

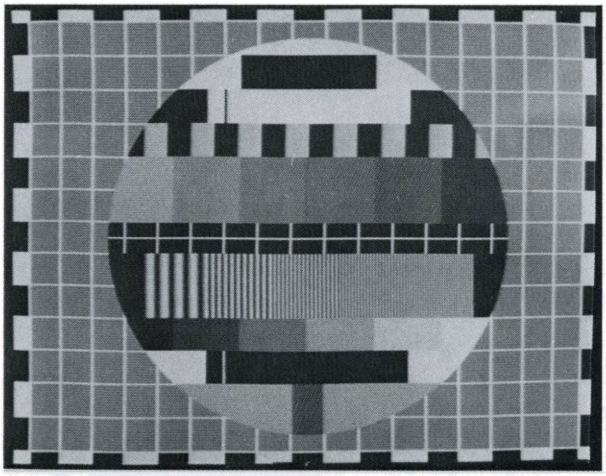
INTERFERING SOURCE	RELATIVE DIRECTION	RELATIVE LEVEL	OFFSET
Kilvey Hill	134°	−33 dB	0
Mynydd Machen	−114°	−42 dB	—

gather some subjective impressions as to how effective a comb filter would be. St. Hilary was already monitoring Stockland Hill at that time as this is now a requirement for the St. Hilary Regional Operations Centre, although the ROC had not then been built.

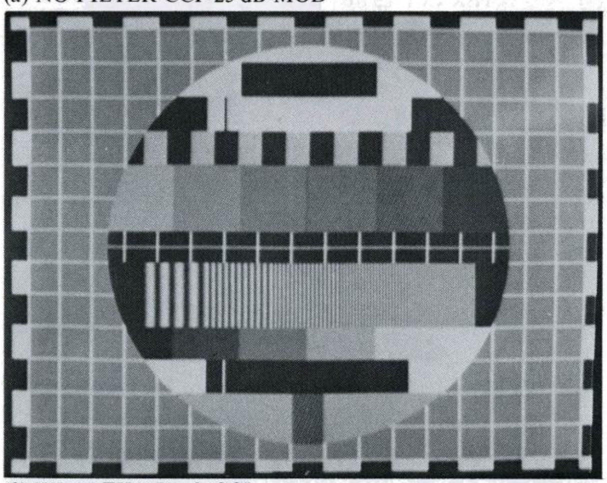
Table 1 shows some of the main interfering signals present on channel 23. Notice that one signal at a single offset predominates; this was from the Kilvey Hill relay transmitter near Swansea. Tests were made on the picture quality as observed on a colour monitor by a team of five or six experienced station engineers. Both modes of the comb filter were used. For each case the comb filter performance was judged and the actual CCI level was recorded.

Results

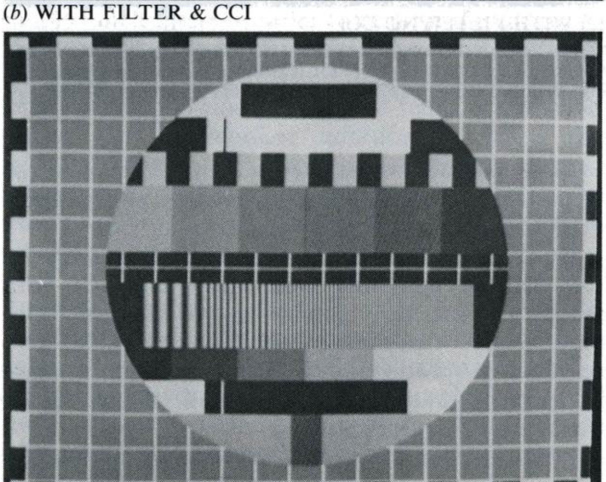
The laboratory test results using the narrow null, can be seen in Fig. 4. It can be observed that the penalty for removing the CCI is a reduction in the wanted picture vertical resolution. This can be seen by observing the reduced sharpness of the horizontals on the test picture. Such degradation is due to the averaging of three lines in a field (i.e. five lines in an interlaced frame). The pulse-and-bar waveform very clearly shows the CCI present in the lower trace and how, in the upper trace, it is removed by the comb filter. It can also be seen how little distortion is produced for a line-repetitive signal: (non line-repetitive signals will be averaged over three lines). For the case of the broad null, (see Fig. 5), a similar level of CCI reduction is shown but this time the picture impairment produced by the filter is only just noticeable. It would be even less so on a less critical, moving, off-air picture. The vertical smearing is reduced by the lowering of the in-phase taps but at the expense of reducing the signal-to-noise ratio as



(*a*) NO FILTER CCI–25 dB MOD



(*b*) WITH FILTER & CCI



(*c*) NO CCI WITH FILTER

explained in Reference 1 for the preceding Section. The waveform shows that the CCI has been greatly reduced and that only a small amount of video distortion is produced (mainly due to residual delay errors in the equipment).

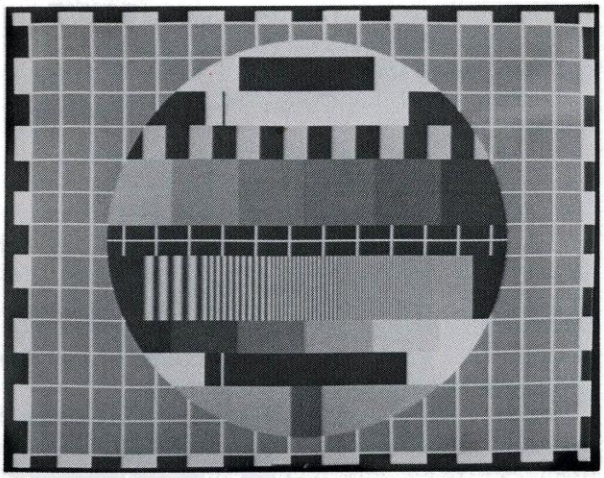
At St. Hilary, the field trials confirmed the laboratory results. With a measured level of CCI at −35 dB (approximately), the comb filter, in both modes, greatly reduced the CCI present. In the narrow null mode the picture improvement was judged to be 1 to $1\frac{1}{2}$ grades on the CCIR 5-point scale. The filter alone appeared to degrade (judged with CCI present) the wanted signal to Grade 3 on the impairment scale (i.e. slightly annoying).

This was due mainly to the loss of some vertical resolution, especially on captions and facial detail when viewed at close range. Using the broad-null mode, the corresponding picture grade improvement was higher at two grades, and the filter degradation lower, Grade 4–$4\frac{1}{2}$ (i.e. just perceptible but not annoying). It was found, in trying to reduce the CCI when using the filter in its broad-null mode, that some difficulty occurred in obtaining a deep null. This could be due to some small level of CCI present at another offset. (The broad-null mode has, unlike the narrow null, only one notch and thus cannot cope with two incoming CCI signals with different offsets.) In addition, the five-control system made optimising the filter setting somewhat complex. Another problem, encountered only in the field trials, was that some monitors do not lock so easily to the field group signal. It must be remembered that the comb filter tries to average the field group over three lines and thus it is distorted. This problem can be overcome by feeding the monitor with external syncs, by using the unfiltered video signal, or by providing the comb filter with an internal sync generator.

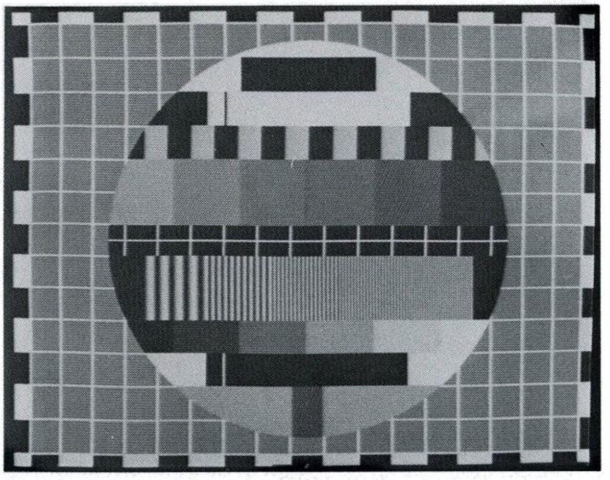
Future Work

The field trials showed that to make the equipment viable as a production device, the controls need to be simplified and reduced in number. One way of achieving this is to use the simplified comb filter shown in Fig. 6. This circuit has been tried

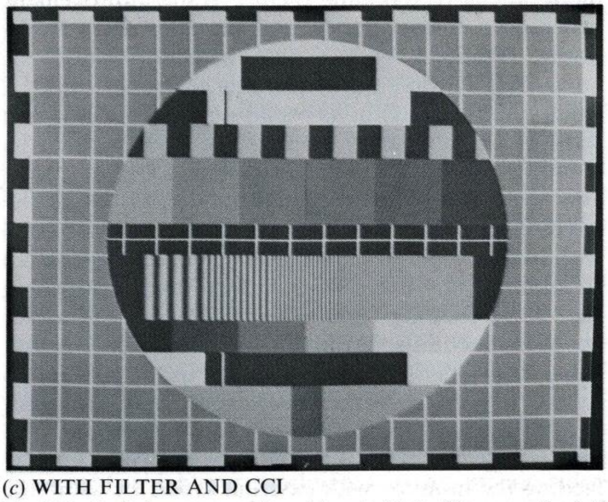
Fig. 4. Typical laboratory results achieved with the filter operating in the narrow-null mode. It can be observed that the penalty paid for eliminating the visible CCI is a just detectable reduction of vertical resolution. This loss of vertical resolution is seen most clearly in the reduced sharpness of the horizontal lines on the test picture and results from the averaging of three lines in a field (five lines in an interlaced frame). The presence and subsequent elimination of CCI can be detected in the pulse and bar waveforms. These results were confirmed in the St. Hilary field trials.



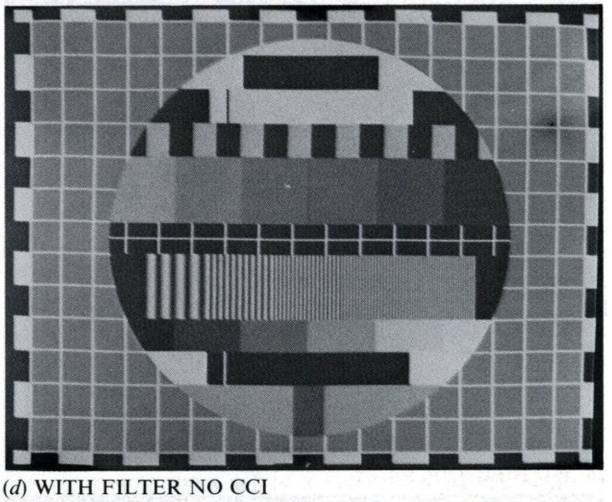
(*a*) WANTED PICTURE NO CCI



(*b*) NO FILTER CCI–25 dB



(*c*) WITH FILTER AND CCI



(*d*) WITH FILTER NO CCI

Fig. 5. Typical laboratory results achieved with the filter operating in the broad-null mode. In this mode a similar reduction in visible interference is achieved as in the narrow-null mode but this time the picture impairment due to loss of vertical resolution is barely noticeable (and would be even less noticeable on 'moving' programme material). The vertical smearing is reduced at the cost of some reduction of the signal-to-noise ratio. These results were confirmed in the St. Hilary field trials.

successfully and provides a one-knob control (marked B). The frequency response, showing a single null, is given in Fig. 7 for both single and double offsets. Present equipment uses glass delay lines to provide the 64 μs video delay but this has several disadvantages. The main drawbacks are: the insertion loss varies with temperature; the need to modulate the video onto an IF carrier; and the high cost of these devices. Future equipment could use charge-coupled-device (CCD) delay lines which are commercially available and which have been tried experimentally. With a reduction in video analogue-to-digital converter costs, it would become practicable to implement the whole filter in digital form, with the advantages of stability and of having a shift register or RAM store as the delay elements.

Nevertheless from the tests reported here it can be seen that the comb filter works effectively under both laboratory and field conditions. Of the two filter modes, the broadband-mode is preferred for two reasons; it produces only a very small picture impairment and the broad null allows some transmitter frequency drift.

It must be appreciated that although comb filtering is very effective for removing CCI from a picture, it cannot eliminate the field group, high-frequency and

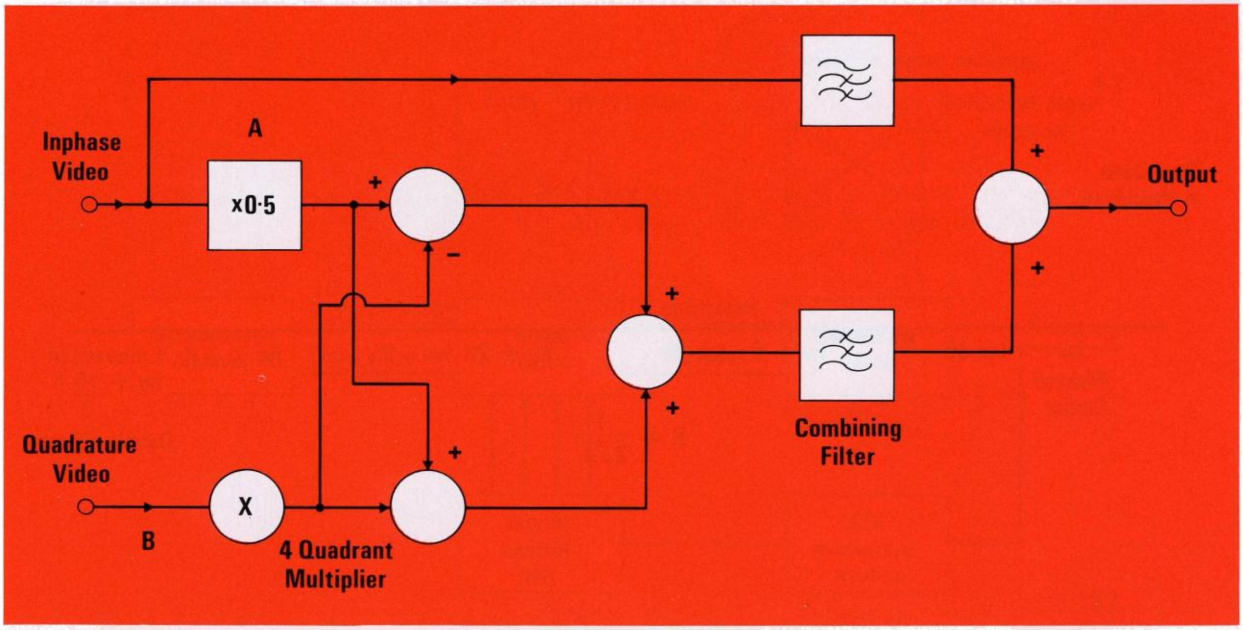


Fig. 6. Design of a simple one-line delay comb filter with one-knob control (adjustment of quadrature video at B). This arrangement has been tried successfully but further refinement is needed for an operational unit. It is also becoming economically viable to consider implementing the entire filter in digital form.

highly-saturated colour information that may be present on the unwanted interference signals. In practice this is not a real problem at realistic levels of CCI (say 20 dB below the wanted signal) or with interfering pictures that do not contain the sort of fine detail that may be present in captions or highly-saturated areas of colour.

There are competing methods of CCI reduction. For example, others have investigated[4] the use of very simple notch filters in order to remove the carrier of the interfering signal. It is clear, however, that this technique will tend to produce somewhat large picture impairments and does nothing to remove unwanted picture line harmonics; additionally, since the rejection notches are necessarily deep and narrow, any drift in the filter notch frequency or in the transmitter frequencies will reduce the amount of CCI rejection.

Adaptive Controls

So far in this section we have been concerned with comb filtering that does not incorporate adaptive control. However it is clear that a more elegant system can be designed around an adaptive or closed control loop, and would be capable of reducing CCI on both of two off-set frequencies.

To close the control loop we require:

(i) An interference measuring system.
(ii) A multivariable control system.

CCI measurement may be achieved by sampling the video signal during the synchronising pulse. During this interval the wanted signal is at a constant, known level.

The control of the comb filter involves more than one variable. In principle there are only two variables '*a*' and '*b*' (see Fig. 3 of preceding Section) due to the symmetry of the system. In practice, it is more convenient to deal with four variables than to try and attain the degree of balance and tracking that would be required for a two-variable control.

The function of the control system is continuously to correct the comb filter coefficients in order to minimise measured interference. This is achieved by optimising the coefficients in sequence, using a perturbation technique to find the setting for minimum interference. The technique is used to control the adaptive aerial array described in an earlier Section.

A complete system block diagram is shown in Fig. 8. The receiver is an IBA re-broadcast receiver designed to operate in the presence of co-channel interference. In-phase and in-quadrature video signals are fed to the comb filter which provides two outputs—filtered and unfiltered, the latter being

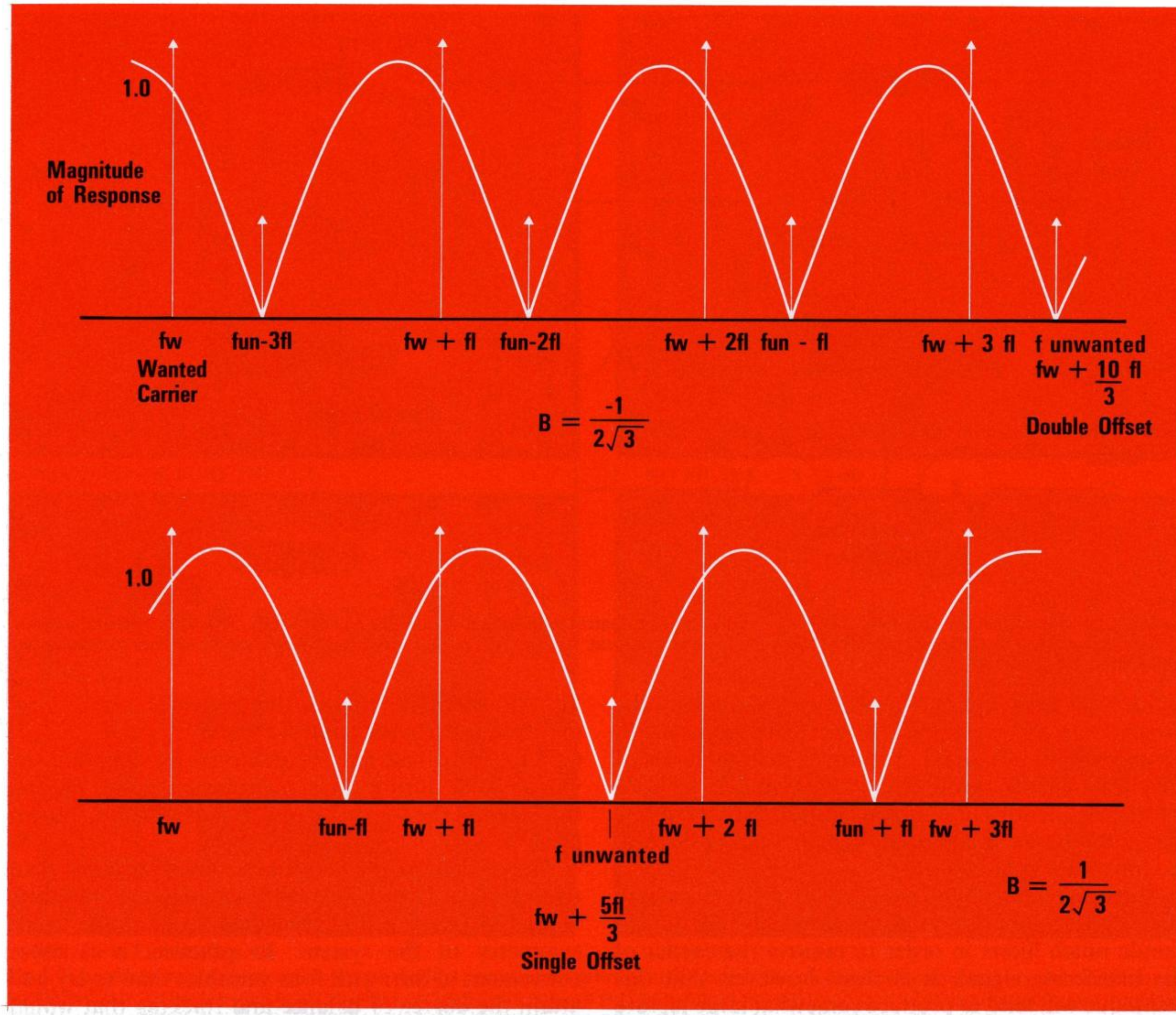


Fig. 7. Response of the simple one-knob filter outlined in Fig. 6 in respect of double and single frequency offsets.

derived from the centre tap. These signals are fed to the combining filter to produce the final video output. Interference is measured on this final output and an error signal fed to the digital control unit. This in turn controls the comb filter, thus closing the feedback loop.

Operation During Field Blanking

A comb filter produces a subjective improvement of the wanted image by summing signals which are advanced or delayed by one line period. During the field blanking interval this process has damaging effects:

(i) The equalising and field synchronising pulse waveforms become distorted.

(ii) Teletext and insertion test signals become corrupted with information from adjacent lines.

For a visual monitoring application it is proposed to provide a separate feed of composite synchronising pulses to the monitor. The re-broadcast application requires a different approach and there appear to be two solutions:

(i) To accept that no CCI cancellation is possible during field blanking; a fast video switch would be used to commutate between filtered and unfiltered video signals in order to inhibit the

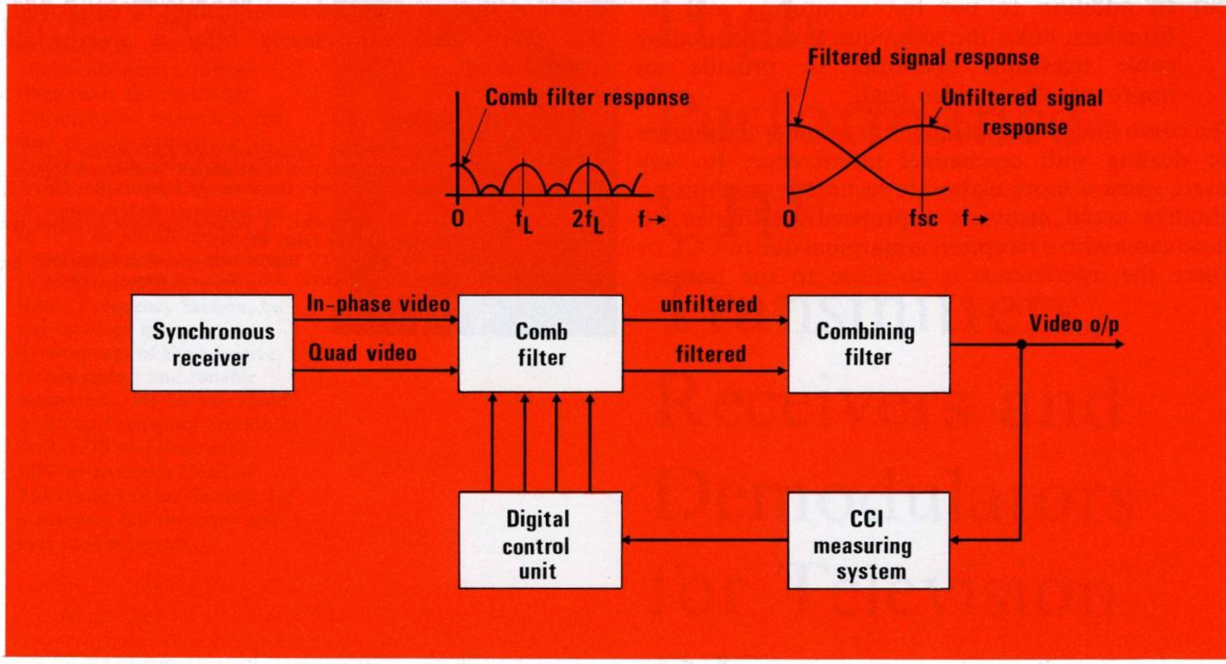


Fig. 8. Block diagram of the experimental comb filter system.

comb-filtered signal during the field-blanking interval.

(ii) To insert locally generated synchronising waveforms and insertion test signals. The teletext signal would be separately decoded, using appropriate techniques to discriminate against CCI, and a regenerated teletext signal inserted.

In the majority of re-broadcast sites for which the comb filter is intended, the interference will be visible on an uncorrected picture, but will rise above the threshold for decoding teletext only for a very small percentage of the time. Consequently option (i) would be adequate.

Conclusions

(i) Comb filtering is an effective way of discriminating against offset co-channel interference and is technically feasible.

(ii) Typically 12 dB of improvement in co-channel protection ratio is possible.

(iii) This improvement may be achieved with a fixed filter in the presence of a single source of interference.

(iv) Where sources are present on both offsets, closed-loop adaptive control is required.

(v) The comb filtering process is subject to constraints both in the frequency and time domains:

(*a*) Comb filtering must be restricted to the lower part of the video band.

(*b*) Comb filtering must be inhibited during the field blanking interval.

(vi) In the presence of one predominant source of interference, there is a trade-off between signal-to-noise ratio and vertical resolution. A complex filter employing in-phase and in-quadrature signals can give optimum results.

(vii) The filter will deal with CCI on both offsets simultaneously, but with a deterioration in vertical resolution.

(viii) Existing re-broadcast receivers of IBA design may be used with a minimum of modification to interface with the comb filter.

(ix) The complete receiving system is self-contained with no changes necessary to aerial and feeder arrangements. This has advantages where the aerial and feeder are shared between more than one broadcasting organisation.

(x) The use of comb filtering can often permit satisfactory UHF reception at sites that would otherwise require installation of a special microwave link, at considerable cost, to provide a programme feed.

(xi) In addition to use for monitoring and re-broadcast links, the technique could be used by cable television operators to provide an improved off-air video feed.

The comb filter is complementary to other techniques for dealing with co-channel interference. In very severe cases a more elaborate technique, such as an adaptive aerial array, is appropriate. However, in those cases where reception is marginal due to CCI, or where the interference is so close to the boresite direction that an adaptive aerial cannot be employed, the comb filter can clearly offer a worthwhile improvement.

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MICHAEL WINDRAM, MA, Ph.D., C.Eng., MIEE, graduated in 1966 and afterwards spent a further three years at Cambridge University on research in the field of radio astronomy. In 1969 he joined Marconi Elliott Avionic Systems Ltd. to work on radar system development. In 1971 he joined the IBA's Experimental & Development Department where, within Radio Frequency Section, he was involved in the development of the adaptive aerials system and tunable equipments. He became Head of Radio Frequency Section in early 1978 and since early 1982 he has been Head of Video and Colour Section. He is married, has two sons and lives in Winchester.

High Performance UHF Transmitters, Receivers and Demodulators for Television RF Measurements

by M. D. Windram

Synopsis

The large number of unattended UHF transmitters built and operated by the IBA created a requirement for a new range of high-quality test transmitters and demodulating equipment with which to carry out the precision measurements associated with assessing, commissioning and maintaining transmitters and RBR receivers. To meet this requirement the IBA developed three tunable equipments: UHF receiver; demodulator; and low-power test transmitter. Specification, design and application of these three sophisticated test equipments are described. A new equipment currently being designed for use by mobile maintenance teams is introduced. Particular features of this comprehensive instrument are its light weight and the use of microprocessor control. The paper also spotlights the practical significance of the phase noise associated with frequency synthesizers, particularly in the presence of incidental phase modulation. This is discussed in more detail in the succeeding Section.

Introduction

The total number of television transmitters operated by the Independent Broadcasting Authority is at present around 700 and is expected to rise to almost 2,000 by the completion of the broadcasting networks. The problems associated with assessing, commissioning and servicing transmitting and monitoring equipment on multiple channels increase in proportion to the number of channels. With this in mind, the IBA has developed a range of high-quality precision UHF test transmitters and demodulating equipment for performance measurement of transmission and reception equipment within the UK Independent Television, Channel Four and Sianel 4 Cymru networks.

A major requirement of the design was to eliminate the need to change costly and heavy fixed channel modules within the test equipment and make the equipment switch or remotely tunable to all channels in the UHF band. A further requirement was that the demodulation equipment should use synchronous detection as first introduced by the IBA to the UHF network for re-broadcast receivers in 1971 because of the superior waveform performance, and that the

modulation equipment should be compatible with equipment using synchronous detection.

To achieve the tunable characteristics required, it was necessary to use a synthesised local oscillator in each of the equipments. The performance requirements of synthesisers when used with synchronous detectors is described in later sections of this paper, and is also of considerable importance to the specification and performance measurement of both synthesisers and synchronous detectors when used elsewhere in a broadcasting network.

The Requirements

There are three main requirements for tunable test equipment within the IBA. These are:

(*a*) Service planning: the tunable receiver is required for assessment of the signal reception at potential transmitter or transposer sites where the off-air signal is required for re-transmission. The tunable test transmitter is used in conjunction with a 1 watt amplifier and a portable tower and aerial to help determine the service area of proposed low-power transposers. This is particularly useful because of the need deliberately to limit coverage to the required area and to avoid interference in other areas.

(*b*) Installation and commissioning: the tunable demodulator and the tunable test transmitter are required for the testing of transmitters and transposers during acceptance, installation and commissioning.

(*c*) Maintenance: the tunable demodulator and tunable test transmitter are required for the maintenance of transmitters and transposers. The tunable receiver is a particularly useful aid in tracking down the origin of distortions etc. in long chains of transmitters and transposers where the cause could lie at any stage in the chain.

Acceptance, commissioning and maintenance limits for transmitters and transposers are shown in Table 1. This list must for reasons of space be limited to only the main parameters.

TABLE 1: ACCEPTANCE AND MAINTENANCE LIMITS

PARAMETER	ACCEPTANCE LIMITS	MAINTENANCE LIMITS
VISION		
$2T$ pulse K rating	2%K	2%K
$2T$ pulse/bar ratio	100% ± 4%	100% ± 4%
bar tilt (10 μs bar)	1%	1%
chrominance/luminance gain ratio	± 10%	± 7%
chrominance/luminance delay	± 20 ns	± 20 ns
chrominance/luminance cross-talk	—	2%
luminance linearity	7%	7%
differential gain (error)	6%	5%
differential phase (error)	3°	4°
intermodulation products (3-tone test)	− 50 dB	− 50 dB
incidental phase modulation	5°	10°
h.f. noise (r.m.s. unweighted)	− 54 dB	− 54 dB
l.f. noise (pk–pk unweighted)	− 50 dB	− 48 dB
SOUND		
noise (weighted—vision modulated)	− 66 dB	− 58 dB

The Equipments

The tunable equipments, the receiver, demodulator and test transmitter are required to be capable of as many of these tests as possible and this indicates the facilities required. The performance limits must also be better than those of the equipment under test. The requirements for a tunable receiver and tunable demodulator were found to be sufficiently different that it was decided not to compromise the performance by producing a single equipment.

Because an off-air signal to be measured or monitored need not necessarily be the strongest signal at that site, the receiver must be capable of operating correctly in the presence of an unwanted group of channels of signal strength higher than that under investigation. Particular features required are therefore low noise figure, high dynamic range, high selectivity and good AGC range.

The tunable demodulator is designed for measurements on transmitter output signals, and therefore very little gain or selectivity is required, but the very high linearity of the receiver must still be maintained. The demodulator is therefore a simplified version of the receiver, but with extra facilities including optional sound rejection to enable transmitter measurements to be made over a wider bandwidth, and with facilities for the measurement of transmitter incidental phase modulation.

The tunable test transmitter is designed to generate a video modulated carrier of adjustable modulation depth from an external video feed. The sound carrier may be modulated either in frequency or amplitude by an internal or external audio source. The equipment also includes facilities for 2- and 3-tone tests with the carrier levels correctly calibrated without adjustment.

All three equipments are required to be rugged as they are for mobile applications and can therefore be expected to be carried by car to remote sites along unmade roads or tracks. Table 2 shows some of the specifications of the three equipments. Again, as for

TABLE 2: TUNABLE EQUIPMENT SPECIFICATION

PARAMETER	RECEIVER	DEMODULATOR	TEST TRANSMITTER
input level	50 μV to 7 mV	100 mV +3, −10 dB 330 mV +3, −10 dB	—
output level	—	—	100 mV ± 1.5 dB
noise figure	8 dB	—	—
VISION PERFORMANCE			
2*T* pulse *K* rating	1% *K*	1% *K*	0.5% *K*
2*T* pulse/bar ratio	100% ± 1%	100% ± 1%	100% ± 1%
bar tilt (10 μs bar)	1%	1%	1%
chrominance/luminance gain ratio	2%	2%	1%
chrominance/luminance delay	± 10 ns	± 10 ns	± 5 ns
chrominance/luminance cross-talk	1%	1%	1%
luminance linearity	1%	1%	0.75%
differential gain (error)	1%	1%	1%
differential phase (error)	1°	1°	1°
intermodulation products	−60 dB	−60 dB	−65 dB
incidental phase modulation	1°	1°	1°
h.f. noise (r.m.s. unweighted)	−50 dB	−53 dB	−52 dB
l.f. noise (pk–pk unweighted)	−56 dB	−56 dB	−54 dB
SOUND PERFORMANCE			
weighted signal/noise ratio (with vision modulated)	60 dB	60 dB	65 dB

Table 1, the complete list would be far too lengthy to include here.

The Designs

The tunable equipments all employ the same design of frequency synthesiser and share a common choice of 1st and 2nd intermediate frequencies. The basic block diagrams are shown in Fig. 1.

The receiver, demodulator and test transmitter are all of double-superheterodyne design with intermediate frequencies of 1191.75 MHz V_c, 1185.75 MHz S_c, 22.5 MHz V_c and 16.5 MHz S_c. Several of the modules of the receiver and demodulator such as the sound trap, vestigial side-band filter, FM demodulator and sound amplifier are also common to the IBA fixed tuned equipments, and will therefore not be discussed here. The synchronous detector, although also common to fixed tuned receivers and demodulators, is discussed here because of the effect on phase noise from synthesisers.

The receiver, demodulator and test transmitter heads are shown in Fig. 2. The image reject and l.o. reject filters which are printed on the various amplifier boards are omitted to simplify the diagrams.

The three RF heads are very similar in design. The receiver and test transmitter use the same 5-pole bandpass filter for the 1191.75 MHz IF. The demodulator, however, does not have the same filtering constraints as the receiver or test transmitter and therefore uses a wider filter bandwidth with consequential easing of alignment and manufacture. A detailed study has been made of the effects of intermodulation and unwanted mixer products. The most complicated case is that of the receiver because of the possibility of many signals being present at the input, with the wanted signal not necessarily the strongest. It was for this reason that the IF of 1191.75 MHz was chosen.

Adjacent channel interference (ACI) is a particularly severe test of a high-performance receiver as intermodulation can occur in stages prior to the ACI filter. It is not possible to design a stable ACI filter at first IF. The early stages of the receiver to second IF prior to the ACI filter are therefore designed with high dynamic range, and the AGC in the earlier stages is limited for this reason.

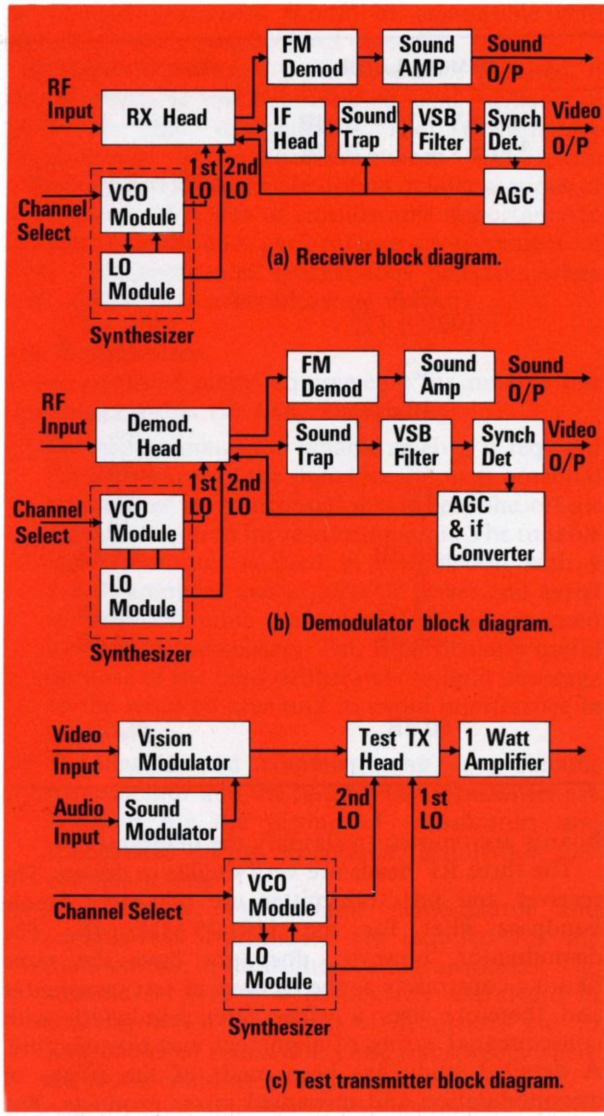


Fig. 1. Block diagrams of (*a*) the tunable receiver; (*b*) the tunable demodulator; and (*c*) the tunable test transmitter. All use the same design of frequency synthesiser and share the same choice of first and second intermediate frequencies (1st IF vision carrier 1191.75 MHz, 2nd IF vision carrier 22.5 MHz). Several of the modules including the synchronous demodulator used in (*a*) and (*b*) are similar to those previously developed for IBA fixed-tuned test equipment.

Figure 3 shows the block diagram of the synthesiser which is common to all three equipments.

The synthesiser frequencies have been chosen to ensure that spurious signals generated in the mixers fall outside the passbands of the IF and RF filters of the three equipments. The phase-locked loop has been optimised to give minimum phase noise on the output signal. In particular, the unusually high reference frequency of 1 MHz was chosen to make possible a high loop bandwidth so that phase noise sidebands out to ~ 50 kHz from carrier for the voltage-controlled oscillator are significantly reduced in amplitude. For all channels, the final peak-to-peak phase noise at the synchronous detector output has been reduced to ~1°. This ensures that the video signal/noise requirements of the three equipments can be met, even for transmissions with high levels of incidental phase modulation. Phase noise is discussed in more detail later. Acquisition of the synthesiser during channel changes take place in less than ~ 100 ms, due to the presence of a powerful frequency lock circuit.

The tunable receiver and demodulator both use synchronous detection for demodulation. This technique has been in use in IBA fixed tuned receivers since 1971 to avoid the quadrature distortion produced as a result of envelope detection of vestigial sideband transmissions.

Figure 4 shows the block diagram of the synchronous detector used for both the receiver and demodulator. The detector phase locked loop has a bandwidth of around 750 Hz. A gated loop is used with the gating at the sync pulse tips, and this sets an upper limit to the stable loop bandwidth available. The use of a gated loop is essential if the detector phase is to get a correct phase reference, essential for both re-broadcast video and for measurement of incidental phase modulation. Non-gated loops are sensitive to modulation and incidental phase modulation and tend therefore to have a varying phase reference.

Figures 5(*a*) and (*b*) show block diagrams of the vision and sound modulators of the tunable test transmitter.

The vision modulator accepts a 1-volt signal and via a peak sync clamp, modulates the 22.5 MHz carrier. Filtering is employed to reduce harmonics on the IF output to a very low level and group delay equalisation is employed to reduce in-band errors to <3 ns. A third tone oscillator is included for 3-tone tests and can be variable in frequency to test for frequency dependence of intermodulation products, and also provides a convenient method of conducting frequency sweeps within the video passband.

The sound modulator generates the 16.5 MHz sound carrier at a spacing of 384 × line rate (6 MHz exactly), or 5.9996 MHz from vision carrier. A narrow-band phase-locked loop is used to provide accurate frequency stabilisation, and can be

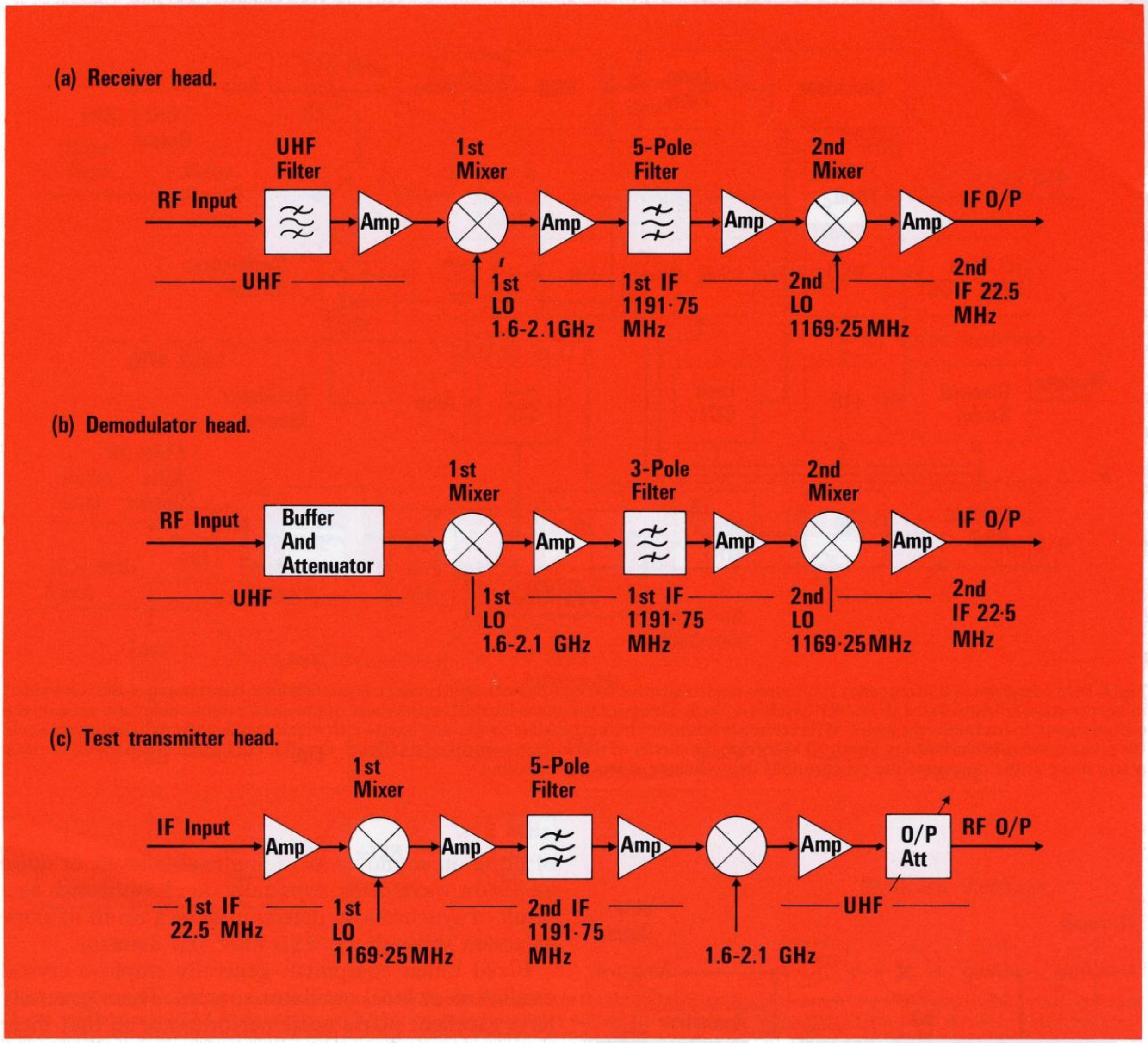


Fig. 2. Simplified block diagrams of the head stages of (*a*) the receiver; (*b*) the demodulator; and (*c*) the test transmitter. The image rejection and local-oscillator rejection filters, which are printed directly on the amplifier boards, are omitted in order to simplify the diagrams. The three head units are basically similar in design and, for example, the same design of five-pole 1191.75 MHz bandpass filter is used in both the receiver and the test transmitter. For the demodulator, where the filtering requirements are different, a filter with wider bandwidth is used to ensure easier alignment and manufacture. For the receiver, effective IF filtering is essential to minimise the effects of intermodulation and unwanted mixer products.

modulated from an audio input with either AM or FM, the latter with switchable 50 µs pre-emphasis. The AM facility is useful for tests of AM rejection of discriminators in receiving and monitoring equipment.

Phase Noise in Synthesisers, Synchronous Detectors and VSB Filters

Phase noise is not generally a directly specified or measured parameter of transmission or measurement equipment. If significant, it manifests itself in the form

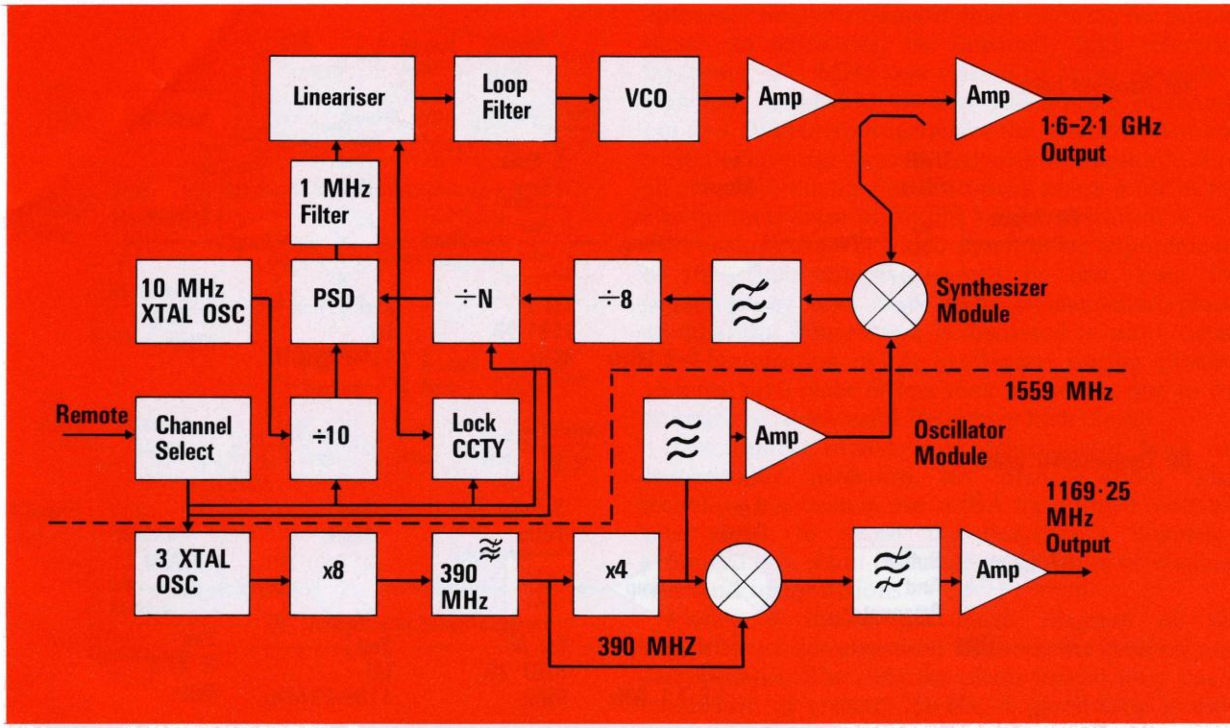


Fig. 3. Block diagram of the frequency synthesiser used in all three test equipments. Synthesiser frequencies have been chosen so that unwanted mixer products fall outside the IF and RF bandpass filters. Design of the phase-locked loop has been optimised for minimum phase noise on the output signal; for example, the high 1 MHz reference frequency makes possible a wide loop bandwidth which significantly reduces in amplitude the phase noise sidebands out to about 50 kHz from the carrier of the voltage-controlled oscillator. On all UHF channels final peak-to-peak phase noise at the output of the synchronous demodulator is less than about 1°.

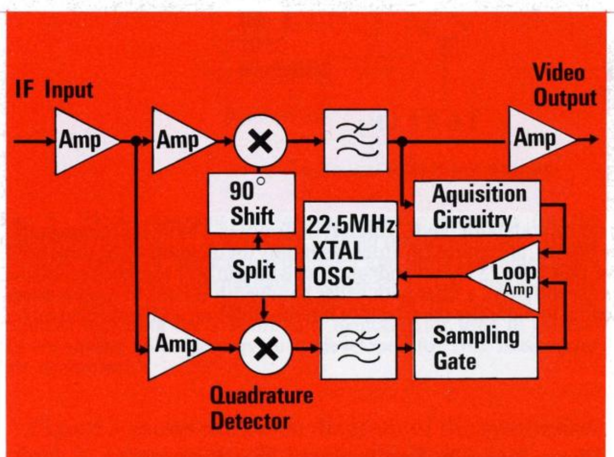


Fig. 4. Block diagram of the synchronous detector used in both the receiver and tunable demodulator. The phase-locked loop has a bandwidth of about 750 Hz. The use of a gated loop is essential to provide a correct phase reference.

of noise on sound or as unpredictable levels of noise on vision waveforms, especially on chrominance as a result of synchronous detection or as a result of slope detection through the VSB filter and detector.

Fixed tuned equipment generally employs crystal oscillators as local oscillator sources. These generally have excellent phase noise performance so that there is no problem for synchronous detection. This is the situation in the existing IBA network where all transmitters use crystal oscillators and multipliers as frequency sources and all permanent demodulators and receivers are fixed tuned.

Synthesisers have considerably higher levels of phase noise in general. Synchronous detectors contain some form of limited bandwidth recovery (a phase locked loop in IBA detectors) and are therefore phase sensitive.

Consider the phase locked loop shown in Fig. 6. θ_ε is the phase error from in-phase in the synchronous detector. The effect of this can be seen in Fig. 7(*a*).

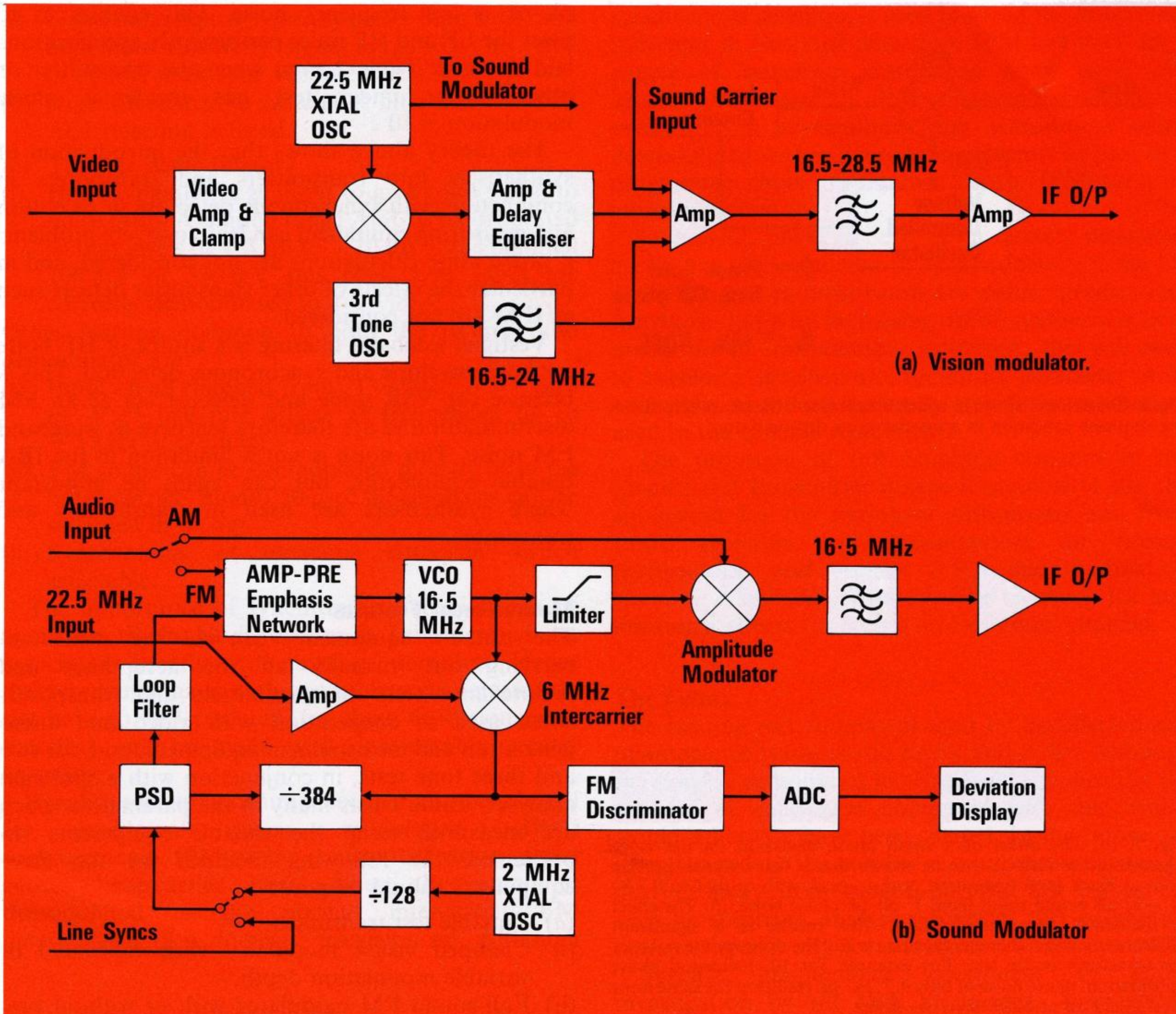


Fig. 5. Block diagrams of (*a*) vision modulator; and (*b*) sound modulator for the tunable test transmitter. A one-volt vision signal, via a peak sync clamp, modulates the 22.5 MHz carrier, with filtering and group delay equalisation to reduce harmonics and in-band errors to a very low level. The sound modulator places the 16.5 MHz sound carrier at 384 times line rate or precisely 5.9996 MHz from the vision carrier. A narrow band phase-locked loop provides accurate frequency stabilisation and the carrier can be amplitude- or frequency-modulated, the latter with switchable 50-microsecond pre-emphasis.

If θ_ε has a total deviation ϕ_ε peak to peak, then the voltage peak to peak at the output of the detector is $V\phi_\varepsilon{}^2/8$ where V is the voltage of the particular part of the waveform of interest with respect to the blanking (zero-carrier) level. For example a video signal-to-noise ratio at black level (measured pk–pk with respect to 0.7 V) of -52 dB is produced by phase noise of 7° pk–pk.

Incidental phase modulation can cause a significant difference between the reference phase and the phase at the appropriate point in the waveform. The detector situation can then be as shown in Fig. 7(*b*). In this case, if $\theta_{IPM} \gg \phi_\varepsilon$, the phase noise pk–pk, then the phase noise is given by $V\phi_\varepsilon \cdot \sin\theta_{IPM}$, so that the signal-to-noise ratio is closely related to the amount of incidental phase modulation. For example, for IPM of 10°, then the same noise level of -52 dB at black level is produced by phase noise of only 0.7°

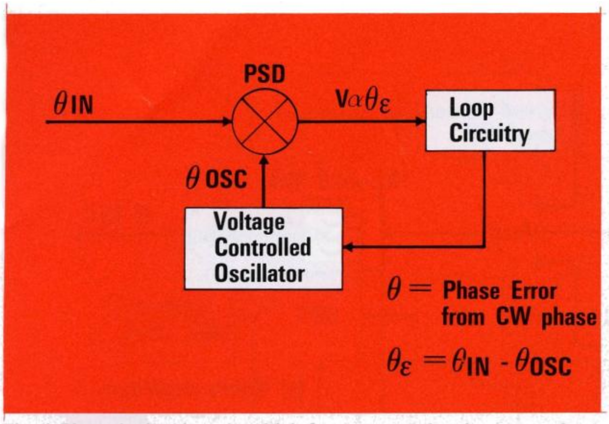


Fig. 6. Phase-locked loop in which θ_ε represents the phase error from the in-phase condition in a synchronous demodulator.

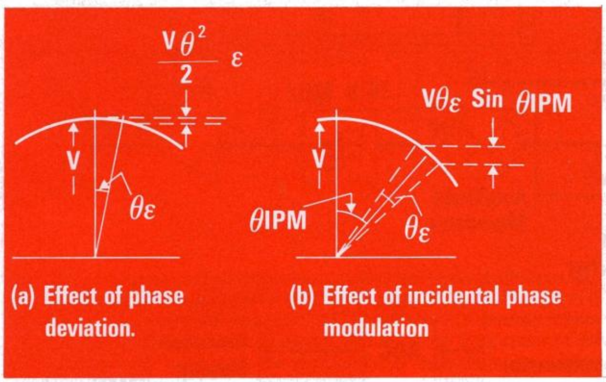


Fig. 7. (*a*) The effect of a small phase deviation on the video signal-to-noise ratio. It can be shown that a video signal-to-noise ratio at black level (measured peak-to-peak with respect to 0.7 V) of −52 dB would result from 7° pk–pk phase noise. (*b*) The effect of incidental phase modulation. This results in a significant difference between the reference phase and the appropriate point of the waveform under test. For example, for 10° incidental phase modulation, phase noise of only 0.7° pk–pk results in the same noise level (−52 dB at black level) as in (*a*).

peak to peak.

For the synthesiser/detector combination used in the tunable equipments described here, care has been taken to minimise the phase noise. In particular, it was necessary to balance the phase noise tracking which results from high synchronous detector loop bandwidth with the aliasing effect of the sampling gate in the detector loop which results in a noise component which increases with increased bandwidth. The optimum has been found to be around 500 Hz–750 Hz loop bandwidth. The total phase noise for each tunable equipment has been reduced to typically ~1° pk–pk of which ~0.5° pk–pk is low frequency noise. This enables us to meet the LF and HF noise performance specifications laid down in Table 2 even when the transmitter or other device under test has incidental phase modulation ~10°.

The theory above shows that the introduction of synthesisers into transmitters and transposers in conjunction with synchronous detectors in reception and measuring equipment can lead to severe problems if phase noise calculations are not considered, and in particular the effects of other transmitter defects such as IPM are not considered.

Vestigial sideband filtering is a source of HF noise for both envelope and synchronous detection. This is because the VSB slope and detector act as an FM discriminator and are therefore sensitive to wideband FM noise. This noise is not a limitation in the IBA tunable equipments, but can again be important where synthesisers are used in transmitters and transposers.

Measurement Features

The tunable equipments provide the means of carrying out virtually all the acceptance and maintenance tests on transmitters and transposers when used in conjunction with video and audio generation and measurement equipment, and, for two and three tone tests, in conjunction with a spectrum analyser. Table 1 lists many of the parameters which are measured using the tunable equipments. In particular, the following facilities for the three equipments should be noted.

(*a*) Tunable test transmitter:

(i) Clamped vision modulator with calibrated or variable modulation depth.

(ii) Full sound FM modulator with or without pre-emphasis. AM facilities are also available for tests on limiters and discriminators. A display of modulation depth is given.

(iii) Switchable 2- or 3-tone tests with calibrated setting of correct carrier levels and fixed or variable frequency third tone.

(iv) Precision output levels and frequencies on all channels from 21 to 70 at −5/3, zero and +5/3 line rate offsets. The channel selection may be made remotely or on the front panel.

(v) In conjunction with the 1 W amplifier, the tunable test transmitter can act as a complete 1 W transmitter for selection of suitable low power transmitter sites.

(*b*) Tunable receiver:

(i) Good noise figure (typically better than 8 dB) and high input sensitivity (signal range 50 μV to 7 mV in 50 rms peak sync vision carrier).
(ii) High selectivity—a switchable filter with >30 dB ACI rejection is fitted.
(*c*) Tunable demodulator:
(i) High dynamic range input to give good signal to noise ratio. A limited range AGC is incorporated so that levels of +3 dB to −10 dB of the nominal 100 mV or 330 mV rms peak sync input are correctly demodulated to 1 V video.

Other features common to both the receiver and demodulator are:
(i) Input on all channels from 21 to 70 at −5/3, zero or +5/3 line rate offsets at frequencies up to ±5 kHz of nominal.
(ii) FM sound can be demodulated directly or in the 'inter-carrier' mode: 50 μsec de-emphasis can be switched in or out.
(iii) Synchronous or envelope detection—switch selectable.

The importance of synchronous detection can be seen from the list of distortions in Table 3 which are associated directly with envelope detection. Hence measurements made with a synchronous detector do not have to be 'corrected' for these distortions, and therefore can be made with considerably greater accuracy.

Synchronous quadrature video can also be selected for incidental phase modulation (IPM) measurements. IPM on the transmitted vision carrier causes 'intercarrier buzz' when the sound is demodulated in the normal intercarrier FM demodulator of a domestic television set. Although synchronous detection is used to avoid the distortions inherrent in envelope detection, IPM can itself be a source of waveform distortions when used with synchronous detection. IPM enhances the noise from oscillators in transmitters and transposers when demodulated with synchronous detectors as part of a programme chain. As shown above, IPM for this must be limited to <10°.

At present, phase noise is not a directly measured parameter. It is assumed to be satisfactory if the l.f. and h.f. signal-to-noise ratios are within specification. It may be that with increasing use of synthesised local oscillators in transmission equipment, this will need to become a specified and measured parameter in its own right, as indeed it already is with the synthesiser used in the tunable equipments.

The provision of full clamping circuitry in the tunable test transmitter makes it possible to use the equipment for the testing of transposers and fixed tuned receivers and demodulators for teletext performance, and because full sound modulation circuitry is included, it is possible to check for any untoward effects occurring on the sound channel.

TABLE 3: DISTORTIONS ASSOCIATED WITH 'PERFECT' ENVELOPE DETECTION—SYSTEM I

	SYNCHRONOUS	ENVELOPE
pulse to bar ratio	100%	~95%
chrominance/luminance cross-talk	0%	~8%
chrominance/luminance gain inequality	0%	~2%
differential gain (140 mV subcarrier)	0%	~4%
line-time non-linearity	0%	~4%

The Future

The existing test equipment used by installation and maintenance teams is still heavy and bulky, although the tunable equipments have reduced significantly the amount of equipment compared with that which would otherwise have been carried. In the future, we intend to reduce the number of equipments carried by moving away from general-purpose test equipment such as spectrum analysers towards test equipment designed specifically for installation or maintenance.

Specific maintenance equipment is under investigation by the IBA at present. A suggested configuration is shown in Fig. 8. The unit being studied at present is the central unit containing all the r.f. processing for all the maintenance tests normally carried out. Microprocessor control will be used to remove the time-consuming setting of controls and calibration of the test equipment. In this way, the total amount of equipment will be reduced and maintenance will be eased, especially for sites with difficult access.

Conclusions

The tunable equipments, the test transmitter, receiver and demodulator, provide the means of carrying out much of the transmitter installation and maintenance work of the IBA broadcast network. These

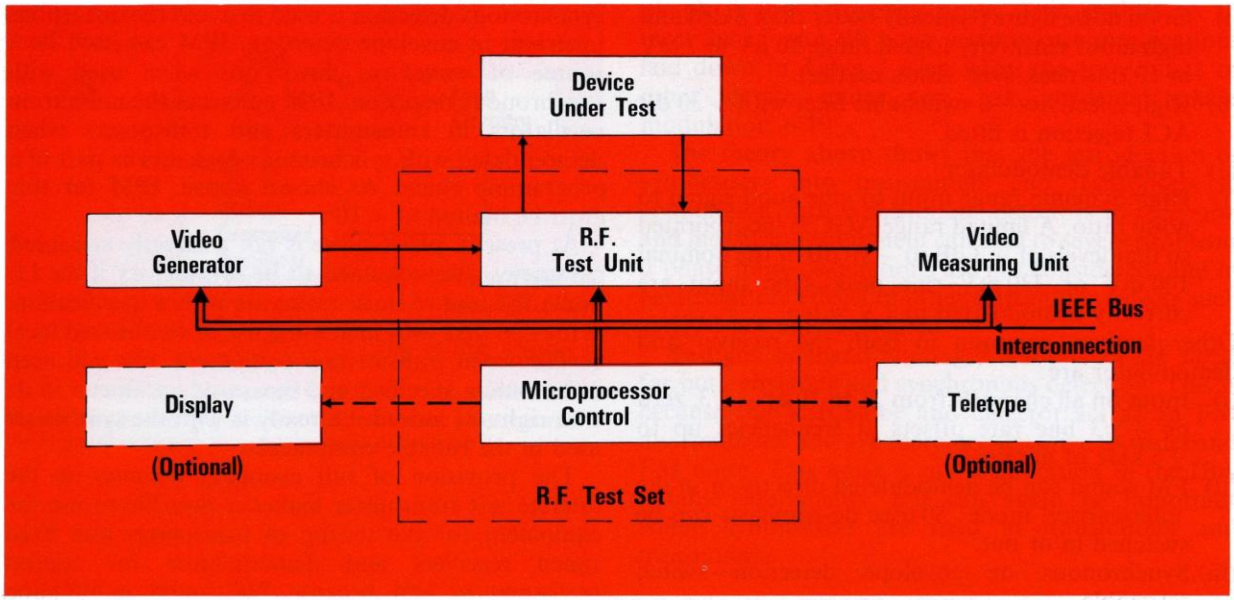
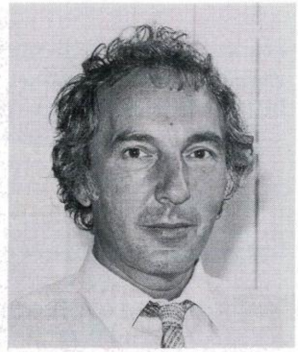


Fig. 8. Comprehensive lightweight transmitter/transposer maintenance test set currently being developed for use by IBA mobile maintenance teams. These will use the latest versions of the tunable test equipment and incorporating microprocessor control.

equipments use low noise frequency synthesisers, and synchronous detectors to provide the high standard of performance required.

In this chapter, the significance of phase noise of synthesisers where synchronous detection is employed later, either in the equipment or in the transmission chain, has been discussed. In particular, the effects of incidental phase modulation have been demonstrated. From this, it can be seen that the phase noise performance of synthesisers used in transmission equipment must be good if noise on demodulated video is not to be a problem. The derivation of a specification is, however, not a simple exercise because of its close dependence on the type and performance of synchronous detector used.

For the future, the tunable equipments will be improved, condensed and automated for a comprehensive lightweight test set for operational use.



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TERRY LONG received his early training at EMI Research Laboratories prior to joining the General Dynamics Corporation in California where he was engaged on the development of receiver and control systems for ship-to-air guided weapons. Whilst in the USA he completed a programme of post-graduate studies at UCLA. Following his return to the United Kingdom in the mid-1960s he joined GEC Electronics at Stanmore where he led the group which developed the tracking receiver for the Rapier defence system. He joined the IBA in 1970 and was appointed Head of Radio Frequency Section in 1973. He has been Head of Experimental and Development Department since 1978.

MICHAEL WINDRAM MA, Ph.D., C.Eng, MIEE. A biographical note appears on page 41.

Synthesiser Phase Noise and its Effect in Broadcasting Systems

by R. J. Barnett, T. J. Long and M. D. Windram

Synopsis

The increasing use of frequency-synthesisers in television transmission equipment, RBR receivers and even some domestic receivers has many advantages. Unfortunately, the spectral purity of the output from a phase-locked loop synthesiser is significantly lower than that of a crystal oscillator. The much higher phase noise has practical effects that should not be ignored; the phase noise of a local oscillator is transferred in a mixer stage, even of the doubly-balanced form, directly on to the signal and consequently degrades system signal-to-noise performance.

The paper provides a brief review of the theory of the phase-locked-loop as used both in UHF frequency synthesisers and in synchronous demodulators, showing the effects of phase noise at various points within the loop. This leads to consideration of the practical design of low-noise loops, and the mechanisms of the conversion process whereby phase noise appears as noise on audio and video signals. The effects of synthesiser noise in a broadcast network are discussed. For example, to achieve a signal-to-noise performance acceptable to broadcast transmission results in a target specification for phase noise in the range 40 Hz to 5.5 MHz of 1 degree peak-to-peak. It is shown that for domestic receivers, based on low-cost components, the effects of phase noise may be far more severe than for broadcast transmission and receiver designers must be aware of the interaction between phase noise, incidental phase modulation and synchronous demodulators.

Frequency synthesisers are beginning to replace fixed-tuned crystal-oscillators for the local oscillators of broadcast transmission and reception equipment. Synthesisers have the advantages of flexibility of tuning and accuracy of frequency but, unfortunately, impose the penalty of higher phase noise. This can be serious and must not be ignored. Phase noise is the noise-like variation of phase of an oscillator output that distinguishes it from an ideal oscillator, or source.

The phase-noise performance of existing crystal-controlled local oscillators is adequate in most cases for broadcast use. A synthesised oscillator will invariably produce a larger amount of phase noise. Any phase noise on a local oscillator in the transmission or reception equipment will be transferred directly onto the signal, and will degrade the overall signal-to-noise performance of the system. The degradation produced will depend on the type of modulation; the effect on frequency modulated (FM) sound will differ from that on amplitude modulated (AM) vestigial sideband (VSB) vision in the television system. The design of the demodulator will also have an important effect on the degree of degradation caused by the phase noise. In particular, the effect of a phase-locked-loop on various noise sources has to be considered in the design of a practical synthesiser.

The block diagram of a phase-locked oscillator is given in Fig. 1, with three sources of noise represented by noise added into the loop at the appropriate points. These noise sources are those of: (1) voltage-controlled oscillator (VCO); (2) the phase-sensitive detector; and (3) reference oscillator noise.

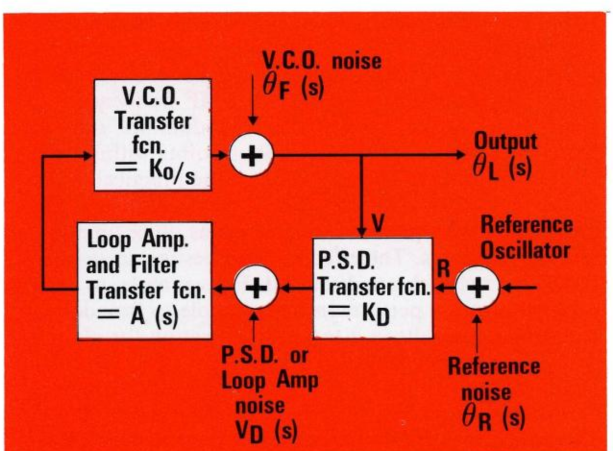

Fig. 1. Block diagram of a phase-locked loop oscillator. The three main sources of noise are voltage-controlled-oscillator noise; phase-sensitive-detector noise; and reference-oscillator noise.

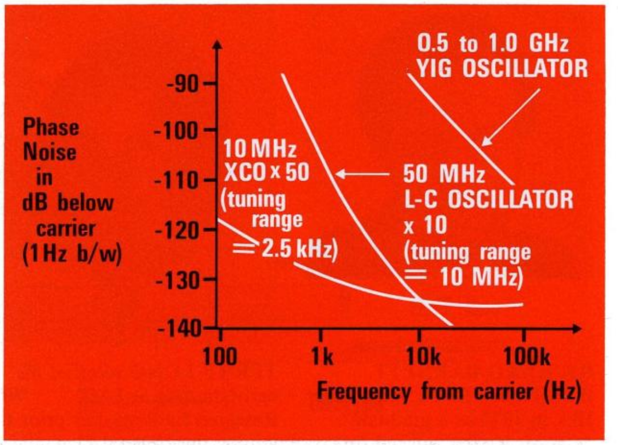

Fig. 2. Comparison of the typical phase-noise sidebands in decibels below carrier (in one hertz bandwidth) for three types of 500 MHz sources: 10 MHz crystal-oscillator with 50-times frequency multiplication; 50 MHz free-running *LC* oscillator with ten-times frequency multiplication; and YIG oscillator on its fundamental frequency. The relation of phase noise to tuning range is clearly shown.

Voltage-controlled Oscillator (VCO) Noise

All oscillators exhibit phase-noise sidebands whose spectral density depends primarily on the *Q*-factor of the oscillator circuit. High-*Q* oscillators, such as crystal oscillators, are unfortunately only capable of being tuned over a very narrow frequency range. The type of oscillator used in a frequency synthesiser is usually required to tune over a much wider frequency range and unfortunately has a much lower *Q* and consequently higher levels of phase noise sidebands. Figure 2 shows a comparison of the phase noise sidebands of different types of oscillator, clearly illustrating the relation of phase noise to the tuning range of the oscillator.

The amplitude of the phase-noise sidebands can be related to the phase deviation of the oscillator by referring to the vector diagram in Fig. 3.

The carrier of amplitude e_c is being phase modulated to give a deviation of θ, resulting in two contra-rotating sidebands, each of amplitude e_s. The peak deviation, θ_{pk} (rads) can be written:

$$\theta_{pk} = \tan^{-1}\left(\frac{2\,e_s}{e_c}\right) \quad \ldots \quad (1)$$

Therefore

$$\theta_{pk} \approx \frac{2\,e_s}{e_c} \quad \text{if } \theta \text{ is small} \quad \ldots \quad (2)$$

Hence the sideband level is directly proportional to

the phase deviation.

To obtain the output response of the loop shown in Fig. 1, for VCO noise only, using normal feedback theory and Laplace notation, and using radians and volts as the units around the loop:

$$\frac{\theta_L(s)}{\theta_F(s)} = \frac{1}{1+K_D A(s) K_0/s} \quad \ldots \quad (3)$$

where $\theta_L(s)$ is the phase noise deviation of the locked VCO

$\theta_F(s)$ is the phase noise deviation of the free-running VCO

K_D is the gain of the phase detector (Volts/rad)

K_o is the gain of the VCO (rads/sec/volt)

$A(s)$ is the loop filter transfer function.

Assuming the loop filter to be an active type so as to give an ideal second order loop, its transfer function is:

$$A(s) = \frac{T_2 s + 1}{T_1 s} \quad \ldots \quad (4)$$

and the loop natural frequency, ω_n, and damping factor, ζ, can be written:

$$\omega_n^2 = \frac{K_0 K_D}{T_1} \quad \ldots \quad (5)$$

$$\zeta = \frac{T_2 \omega_n}{2} \quad \ldots \quad (6)$$

Now the loop response can be written as:

$$\frac{\theta_L(s)}{\theta_F(s)} = \frac{s^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad \ldots \quad (7)$$

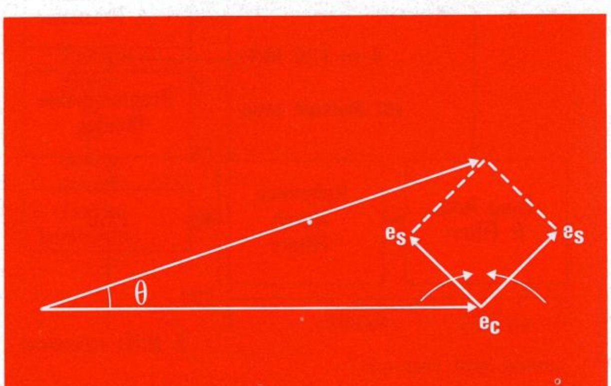

Fig. 3. Vector diagram of a phase (angle) modulated carrier showing how the amplitude of the phase-noise sidebands is related to the phase deviation of the oscillator. For small deviation the sideband level is directly proportional to the phase deviation.

and substituting $s = j\omega$ to give the frequency response:

$$\frac{\theta_L(\omega)}{\theta_F(\omega)} = \frac{\omega^2}{(\omega^2 - \omega_n^2) - j2\zeta\omega_n\omega} \quad \ldots \quad (8)$$

From equation (8) it can be seen that if $\omega \gg \omega_n$, the response is unity and the phase noise sidebands of the locked VCO are the same as the free-running VCO. If $\omega \ll \omega_n$, the phase deviation and hence the phase noise sidebands are suppressed in the ratio $(\omega/\omega_n)^2$.

Phase Sensitive Detector (PSD) Noise

The second source of noise in Fig. 1 is due to the phase sensitive detector or the loop amplifier. Following a similar technique as used to derive equation (8), the output response of the loop to a noise input from the PSD, $V_D(s)$, is:

$$\frac{\theta_L(\omega)}{V_D(\omega)} = \frac{-(\omega_n^2 + j2\zeta\omega_n\omega)/K_D}{(\omega^2 - \omega_n^2) - j2\zeta\omega_n\omega} \quad \ldots \quad (9)$$

Using equation (9), if $\omega \gg \omega_n$, the response becomes:

$$\frac{\theta_L(\omega)}{V_D(\omega)} = \frac{-j2\zeta\omega_n}{K_D\omega} \quad \ldots \quad (10)$$

and the phase sidebands are dependent on the loop parameters and falling at 6 dB/octave away from the carrier.

For $\omega \ll \omega_n$, the response is constant and equal to $1/K_D$.

Reference Oscillator Noise

The final noise source input of Fig. 1 represents phase noise present on the reference oscillator, $\theta_R(\omega)$. To analyse the effect of this noise it is possible to replace the phase noise source, $\theta_R(\omega)$, on the reference input of the PSD by an equal and opposite phase noise source, $-\theta_R(\omega)$, on the input from the VCO. A similar technique to that used before gives:

$$\frac{\theta_L(\omega)}{\theta_R(\omega)} = \frac{(\omega_n^2 + j2\zeta\omega_n\omega)}{(\omega^2 - \omega_n^2) - j2\zeta\omega_n\omega} \quad \ldots \quad (11)$$

This is identical to equation (9) except for the sign and the absence of K_D in the denominator, and the same approximations are true.

Loop Noise

EFFECT OF FREQUENCY DIVIDER IN THE LOOP: It is common practice to include a frequency divider in a frequency synthesiser, between the VCO and PSD. This effectively reduces the PSD gain by the division

ratio N. Any noise from the PSD, loop amplifier or reference oscillator is *increased* by the factor N, making their design more critical.

EFFECT OF DELAY IN THE LOOP: There are various possible causes of delay within the loop; some are due to the absolute delay through the various control circuits, e.g. operational amplifiers, or to some time-constant other than that designed into the loop filter, e.g. reference notch filters or VCO tuning port bandwidth. A more subtle cause of delay can be due to the type of PSD used. If the PSD has digital signals at its inputs, there is an inherent delay which has a mean value of one half of the reference frequency period. If the PSD is of the sample-and-hold type, there is an additional delay, again of one half of the reference frequency period, due to the hold action. The latter two causes of delay can have a serious effect upon the noise performance of the synthesiser and will determine the ratio of maximum allowable loop bandwidth to reference frequency ratio.

Inclusion of a delay in the derivation of equations (8), (9) and (11) is straightforward. However the results cannot be factorised and are best investigated by computer graphics.

EFFECT OF A SAMPLER IN THE LOOP: In many applications the control signal within a phase-locked loop may be sampled periodically. This may apply in the case of a synthesiser where the PSD operation involves sampling at the reference frequency. In this case if the bandwidth of the noise input to the sampler exceeds one half of the sampling frequency, a violation of Nyquist's sampling criterion exists, and there will be aliasing of the high frequency noise into the lower frequency range. In the synthesiser this noise aliasing effect can usually be ignored provided the noise of the VCO at the reference frequency spacing from the carrier is much less than its low frequency value. However, in the case of the synchronous demodulator the effect is important and must be taken into account in any noise analysis.

Synthesiser Design Considerations

The main purpose of the phase-locked loop in a frequency synthesiser is to control the frequency of the VCO at some multiple or fraction of that of a stable reference oscillator. However the loop may serve also to suppress some of the phase noise sidebands of the VCO, as already noted. The phase noise sidebands are reduced in the ratio $(\omega/\omega_n)^2$ within the loop bandwidth, where ω_n is the loop natural frequency or loop bandwidth. Therefore, to maximise the suppression of the VCO phase noise, as wide a loop bandwidth as possible should be chosen.

Unfortunately, however, by increasing the loop bandwidth the contributions to the output noise from the PSD, loop amplifier and reference oscillator are increased. This may be a limiting factor to the allowable loop bandwidth, but careful design of these circuit elements should minimise their noise contribution.

There are other constraints in the design of a frequency synthesiser, depending on channel spacing, speed of operation of dividers, reference frequencies available, etc. These constraints are likely to put an upper limit on the reference frequency used at the phase detector of the phase-locked loop. This limit on the reference frequency partly determines the choice of loop bandwidth. It has been found that the ratio of reference frequency to loop bandwidth must be greater than about 20 to 1, assuming a sample-and-hold type of PSD and depending on the noise of the VCO and the required noise performance of the synthesiser. If this ratio is reduced, the delay around the loop, described earlier, causes peaks in the loop

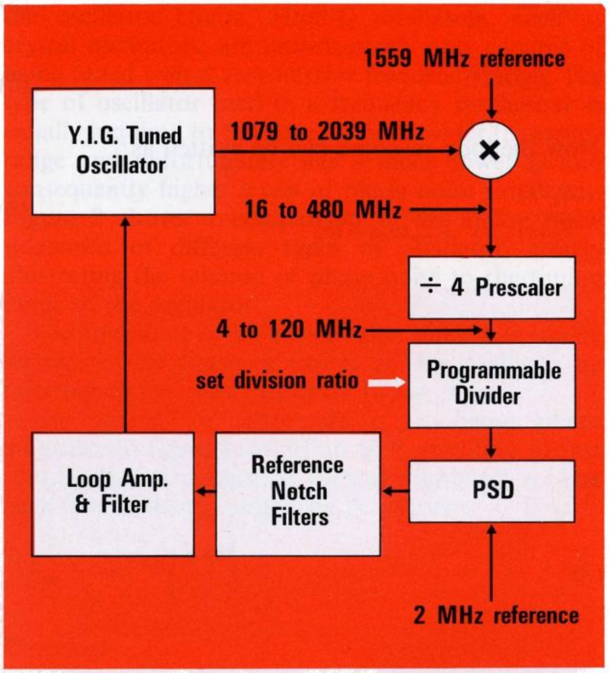

Fig. 4. Block diagram of a frequency synthesiser designed by IBA engineers for use in high-performance measurement equipment. It uses a YIG oscillator with an octave tuning range in a loop of 50 kHz bandwidth with a 2 MHz crystal-controlled reference oscillator.

response at the loop natural frequency and hence excessive noise peaks in the output spectrum of the synthesiser.

The block diagram of Fig. 4 shows a synthesiser designed by the IBA for use in high quality TV measurement equipment. It uses an octave tuning range YIG oscillator in a wideband loop of bandwidth 50 kHz, with a reference frequency of 2 MHz. This is based on a successful design of loop first used in 1974.

The effective delay around the loop amounts to approximately 0.7 μS, causing a 3 dB peak in the noise sidebands at the loop natural frequency. The programmable divider operates with an input signal up to 120 MHz in order to maintain the reference frequency at 2 MHz. The divider is a straightforward divide-by-N counter in order to allow continuous division ratios as low as N equals two.

Using the theory developed earlier, the computer predictions of the synthesiser noise performance were obtained and are shown in Fig. 5. In Fig. 5(a) curve (1) shows the measured free-running, YIG-oscillator noise and curve (2), the result of locking this in a noise free, zero delay loop. Curve (3) shows the effect of including a delay of 0.7 μS in the loop. This is shown again in Fig. 5(b) along with contributions to the output noise from the PSD/divider and reference sources shown in curves (4) and (5) respectively. The total noise, which is the sum of curves (3), (4) and (5), is shown in curve (6) with the phase-noise data measured on the synthesiser shown as ringed data points.

Effects in the Vision Channel

The vision channel of a TV broadcast system is an amplitude modulated (AM), vestigial-sideband (VSB) system. Sensitivity to phase noise added to the signal by the synthesised local oscillators arises in two ways. Phase noise is converted to video noise by the processes discussed below.

NOISE CONVERSION IN THE IF FILTER: The VSB filter used at IF in a receiver has the response shown in Fig. 6. On the vestige slope of this response the phase-noise sidebands on the signal are differentially amplified, according to their spacing from the vision carrier. This asymmetry in the amplitude of the phase-noise sidebands can be resolved into the original phase-modulation (PM) sidebands and additional amplitude-modulated (AM) sidebands of amplitude $f_n/1.25$ times that of the original phase sidebands, where f_n is the frequency spacing from the carrier in MHz of the noise sideband. For small values of f_n, i.e. noise close to the carrier, the PM to AM conversion is minimal. For noise greater than 1.25 MHz from the carrier, the VSB filter removes one phase-noise sideband and attenuates the vision carrier by 6 dB, resulting in an equivalent AM sideband-to-carrier

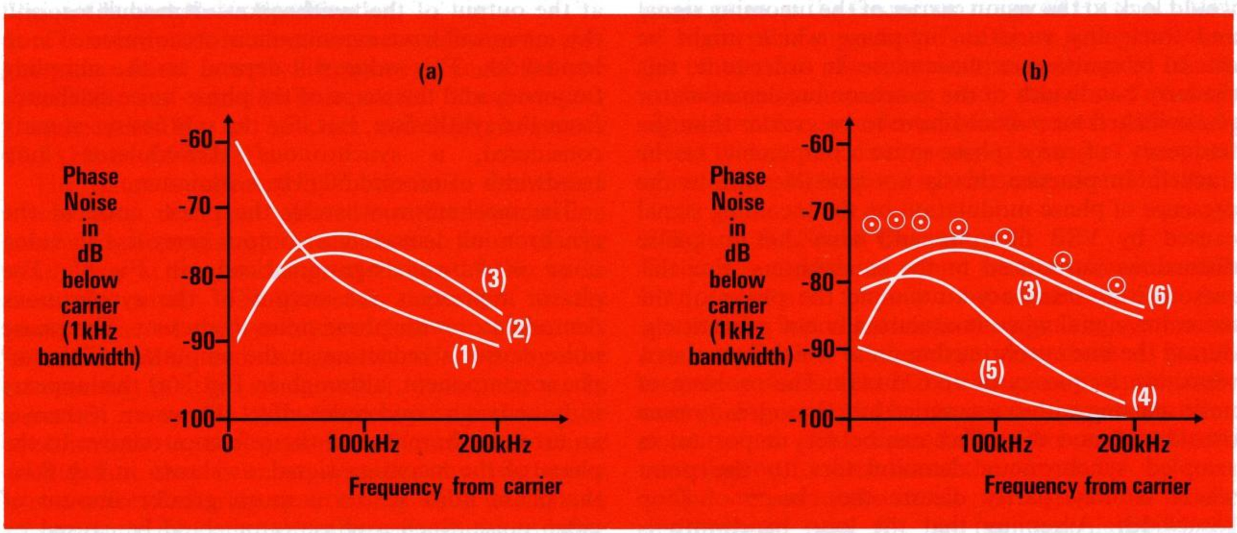

Fig. 5. Computer predictions of the noise contributions to the phase-noise performance of the frequency-synthesiser shown in Fig. 4. Curve (1) relates to free-running YIG oscillator; (2) YIG oscillator locked with noise-free, zero-delay loop; (3) effect of including 0.7 μs delay in the loop. Curve (3) is repeated in (b) where curves (4) and (5) are the noise contributions of the PSD/divider and reference sources respectively. Curve (6) is the total predicted noise output (the sum of curves (3), (4) and (5)) with measured results shown as ringed data points.

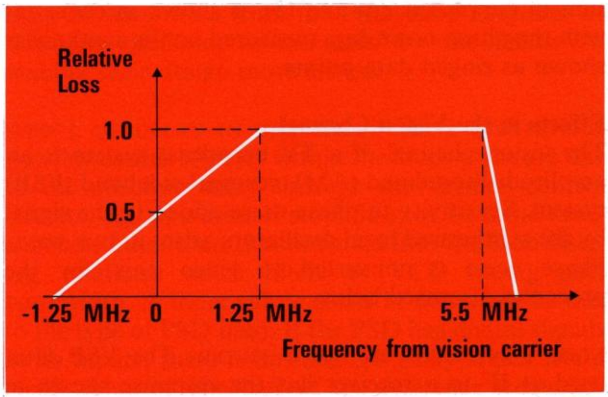

Fig. 6. The vestigial-sideband filter in the receiver is designed to have standard response characteristics. The result of PM to AM conversion in a VSB filter may affect the high-frequency signal-to-noise performance if the voltage-controlled oscillator in the synthesiser has a high noise floor at frequencies extending more than about 1 MHz from the carrier.

ratio equal to the original PM sideband-to-carrier ratio.

In practice the result of the PM to AM conversion in the VSB filter may affect the high-frequency video signal-to-noise performance if the VCO in the synthesiser has a high noise floor at frequencies greater than about 1 MHz from the carrier.

NOISE CONVERSION IN THE SYNCHRONOUS DEMODULATOR: Ideally a synchronous demodulator would lock to the vision carrier of the incoming signal and track any variation in phase which might be caused by synthesiser phase noise. In order to do this the loop bandwidth of the synchronous demodulator phase-locked loop would have to be greater than the frequency of any phase-noise component to be tracked. In practice this is not possible due to the presence of phase-modulation on the incoming signal caused by VSB filtering, and also due to phase distortions introduced in the transmitters. For this reason, it is necessary to sample the phase of the incoming signal when modulation is not present, e.g. during the line synchronising pulse which occurs at a repetition frequency of 15.625 kHz. The problem of noise aliasing within a sampled loop has already been considered, and this effect can be very important in sampled synchronous demodulators to the point where it may partly dictate the choice of loop bandwidth. Assuming that the loop bandwidth is much less than the sampling frequency, the amount of high-frequency noise aliased into the low frequency range, is directly proportional to the loop bandwidth.

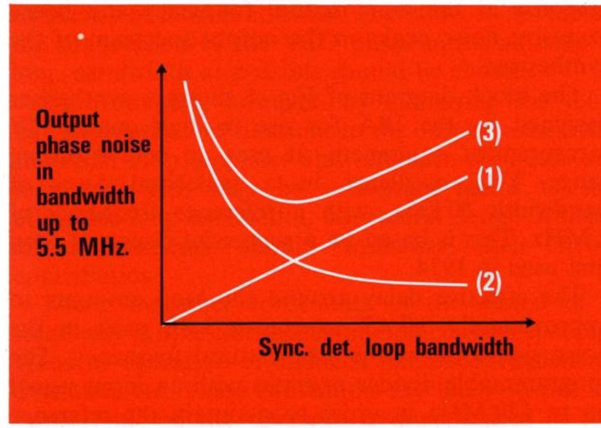

Fig. 7. Phase noise as a function of loop bandwidth at the output of the synchronous demodulator due to various mechanisms. Curve (1) shows noise contribution due to aliasing of high-frequency noise into the low-frequency range; (2) original (non-aliased) noise not tracked out by the synchronous demodulator loop; and (3) sum of these two components present at the output of the synchronous demodulator, and this will have a minimum output at some specific value of loop bandwidth. For the frequency synthesiser discussed earlier in this Section a loop bandwidth of about 400 Hz is optimum.

Figure 7, curve (1), shows this noise contribution due to aliasing as a function of loop bandwidth. Curve (2) shows the original (non-aliased) noise which would not be tracked out by the synchronous demodulator loop as a function also of loop bandwidth. The sum of these two noise components, curve (3), will be present at the output of the synchronous demodulator, and this curve will have a minimum at some value of loop bandwidth. This value will depend on the sampling frequency and the shape of the phase-noise sidebands from the synthesiser, but, for the synthesiser already considered, a synchronous demodulator loop bandwidth of around 500 Hz is optimum.

The mechanism whereby the phase noise at the synchronous demodulator output gives rise to video noise is shown diagrammatically in Fig. 8[2]. The phasor represents the output of the synchronous demodulator with phase noise θ_n present. The phase noise causes a reduction in the amplitude of the in-phase component, although in Fig. 8(a) this appears to be only a second order effect. However, if there is an error in the phase of the reference relative to the phase of the incoming signal, as shown in Fig. 8(b), the phase noise causes a much greater amount of video noise. Such a phase error could be caused by incidental phase modulation on the incoming signal or be due to misalignment in the synchronous demodulator.

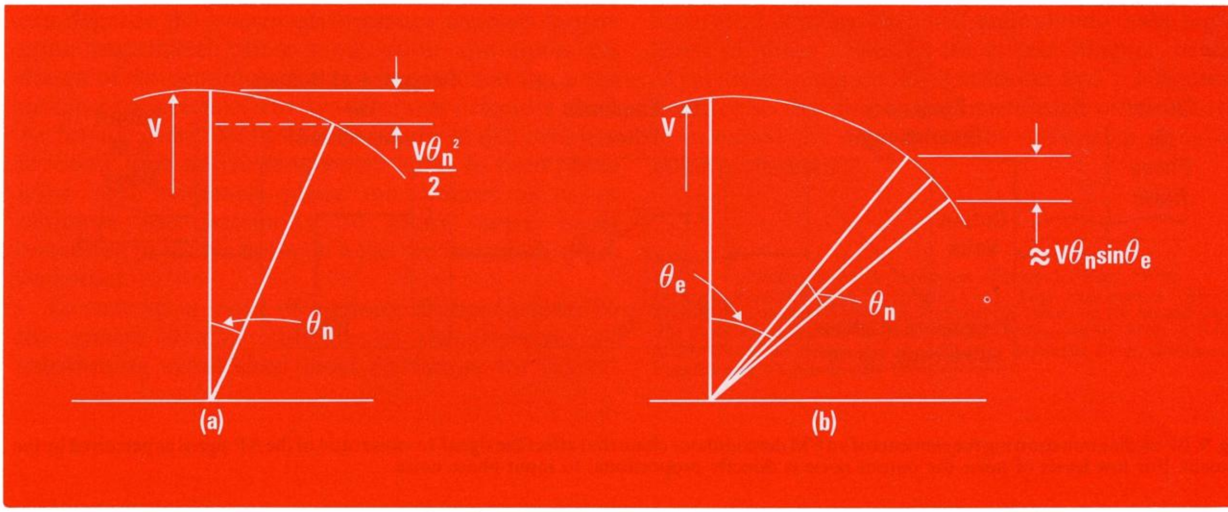

Fig. 8. Phasor diagrams of the output of the synchronous demodulator showing the mechanism whereby phase noise results in a degradation of the video signal-to-noise ratio. As indicated in (b) a phase error in the reference signal relative to the incoming signal is the more serious.

Phase noise is also added to the chrominance sub-carrier and can cause errors in the phase and hence the hue of the signal. Because of the phase switching in PAL colour, the effect of phase noise is different, and in fact considerably less than for NTSC. In PAL the degradation due to phase noise on the chrominance sub-carrier is generally acceptable if other noise parameters are satisfied.

Vision Channel Noise Specification

It is conventional in television measurements to split the video band into two frequency ranges for the specification of signal-to-noise ratios. These are low frequency (40 Hz to 7.5 kHz) and high frequency (7.5 kHz to 5.5 MHz)[3].

The latter is affected by both noise conversion mechanisms described above, but the degradation produced depends upon the noise floor of the synthesiser at frequencies greater than 1 MHz or so from the carrier. The former is dependent on the phase noise of the synthesiser close to carrier, i.e. below 1 MHz, and is greatly affected by the parameters of the synchronous demodulator. In practice it is the low-frequency signal-to-noise specification in the presence of incidental phase modulation that is most degraded by synthesiser phase noise. Referring again to Fig. 8(b), and taking into account the definition of signal-to-noise as being noise relative to picture amplitude, the noise-to-signal ratio, N/S, due to the phase noise conversion in the synchronous detector can be written:

$$N/S = 0.0237\ \theta_n \sin\theta_e. \qquad (12)$$

where θ_n is the phase noise in degrees
and θ_e is the phase error.

As an example of this, for incidental phase modulation of 10° and phase noise of 0.7°, a low-frequency signal-to-noise ratio of 51 dB results. To achieve signal-to-noise performance of an acceptable level for broadcast use[2], a target specification for phase noise in the frequency range 40 Hz to 5.5 MHz of 1° pk–pk is suggested. This figure is dependent on the shape of the phase-noise sidebands of the synthesiser and assumes that the low frequency signal-to-noise specification is the most difficult to meet. It does not assume any tracking of the phase noise above 40 Hz by a synchronous demodulator.

Effects in the Sound Channel

Frequency modulated sound transmission is used in both radio and television broadcast systems. Any phase noise introduced onto the signal by synthesised local oscillators is therefore demodulated by the frequency discriminator to produce noise at the audio output. Figure 9 shows the processing of the phase noise in an FM demodulator to arrive at the contribution to the noise at the output.

The output audio noise as a function of the discriminator gain, K_D, has been calculated for the predicted synthesiser noise response of Fig. 5. Integrating this noise spectrum to obtain the total r.m.s. noise voltage gives a noise ratio when compared

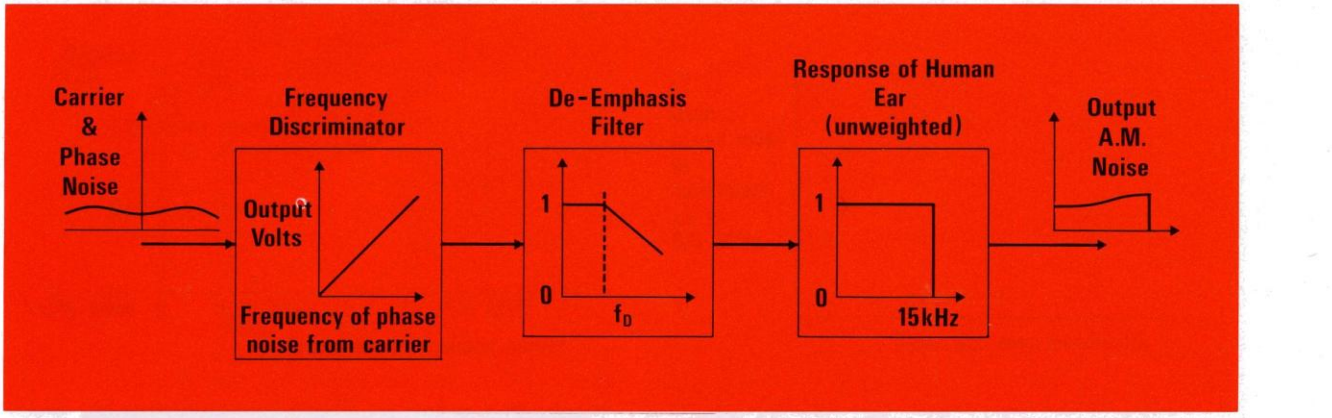

Fig. 9. Block diagram showing the elements of an FM demodulator chain that affect the signal-to-noise ratio of the AF signal as perceived by the listener. For low levels of noise the output noise is directly proportional to input phase noise.

to a signal at a modulation frequency of 400 Hz, and a peak deviation of 50 kHz of −72 dB. The specification calls for a value of −60 dB.

A point worth noting is that, for low levels of noise, there is a one to one correspondence of noise components in the output frequency spectrum for phase noise components at the input to the discriminator. It is only the phase noise in the frequency range from 20 Hz to 15 kHz from the carrier that affects the final output signal-to-noise.

The technique by which the sound carrier is recovered may affect the audio signal-to-noise degradation caused by synthesiser phase noise. If intercarrier techniques are used, and the synthesiser phase noise is added at the transmitter such that the noise on the vision carrier is non-coherent with the sound carrier, then the resulting demodulated noise level may be due to the sum of vision and sound phase noise. If, however, the same synthesiser phase noise is common to both sound and vision carriers, then there may be a cancellation of the phase noise when the sound carrier is obtained by detecting the vision carrier.

Effects on the Broadcast Network

The essential difference between a re-broadcast transmitter and a transposer is that, in the former case, the signal is demodulated to video and, in the latter case, the signal is not demodulated, but merely changed in frequency. With a chain of transmitters, the video noise contribution due to phase noise can be considered for each link separately and added conventionally.

Where transposers are involved, both incidental phase modulation and phase noise accumulate along a chain so that the total signal-to-noise degradation at the receiver at the end of the chain is obtained by applying equation 12, using the accumulated values of θ_n and θ_e. This means that in considering the phase noise specification of a transposer, it is essential to consider not only its own relatively low incidental phase modulation, but also the higher incidental phase modulation of the parent transmitter.

The effects of synthesiser phase noise remain within the control of the broadcasters only until the signal is transmitted. While within their control, it is possible to ensure that phase noise and incidental phase modulation remain within specified limits which ensure no degradation in low or high frequency signal-to-noise ratio of the picture.

However, phase noise can occur in the domestic receiver if synthesised tuners are used. In this case, the effects are likely to be far more severe because the need for low cost prevents the use of very low noise oscillators and loop components. It is therefore essential that the receiver designer be aware of the interaction between phase noise, incidental phase modulation and synchronous detectors.

Conclusions

It has been shown that the television and radio broadcast system is sensitive to phase noise especially when the phase noise increases close to carrier as might be caused by the use of synthesised local oscillator sources. The phase noise can degrade both the low frequency and high frequency video and sound signal-to-noise performance although the most sensitive parameter depending on the synthesiser noise characteristics, is the low frequency video signal-to-noise ratio. The amount by which the phase noise

can degrade the system performance depends largely upon transmitter phase modulation and upon the design of the demodulator, and in particular upon the design parameters of any synchronous detectors used. As broadcasters, we have been aware of this problem for some time and have specified limits to both phase noise and incidental phase modulation so as to minimise the possibility of phase noise being converted to visible noise on a picture or audible noise on sound.

To sum up, a word of warning is appropriate. To the system designer beware of the problems of substituting synthesised local oscillators for crystal controlled sources; they will undoubtedly have higher levels of phase noise. To the circuit designer, beware of the need to design the synthesiser and synchronous detector so that they complement each other in order to minimise the degradation to the system signal-to-noise performance.

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3. *IBA Technical Review Vol. 2*, 'Technical Reference Book' (Independent Broadcasting Authority, July 1974).

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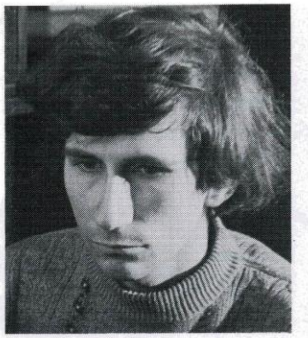


An Improved Demultiplexer for Stereo or Three-channel Broadcasts

by J. Halliday

Synopsis

Following several years' investigation of surround-sound systems and techniques, IBA engineers developed an improved three-channel matrix system designated MSC (patent applied for). This system, which has been used in experimental over-air broadcasts, is the first surround-sound system that is fully compatible subjectively with existing stereo receivers, as described in *IBA Technical Review Vol. 14*.

An important part of this work was aimed at devising low-cost circuitry suitable for use in high-quality domestic receivers. This has resulted in a new and improved demultiplexer that is suitable not only for three-channel surround-sound broadcasts but also as a superior-quality stereo decoder.

The paper identifies common shortcomings in conventional stereo decoders, including the production of spurious 'birdies' (i.e. twittering noises) resulting from adjacent-channel transmissions and degradation of the signal-to-noise ratio due to the phase-noise of phase-locked loops.

The improved demultiplexer features novel demodulators which obviate the need for 'anti-birdie' filtering and a dual-mode phase-locked loop system which meets the more stringent requirements of three-channel reception without the need for critical preset adjustments. Fundamental design theory together with basic circuit details are presented for a reasonably simple design which is capable of good domestic reception of three-channel transmission broadcasts without critical setting-up adjustments and which is also capable of excellent stereo reception.

FM radio broadcasting in Band II is used in many countries to provide a stereophonic (two-channel) programme service, mostly using the pilot-tone multiplexing system[1, 2]. There have now been a number of proposals for broadcasting in surround-sound using one or more additional audio channels (for example, refs. 3–6). In Europe, at least, it seems likely that the only options of immediate interest will be those using just three channels. In all the proposed systems these three channels would be multiplexed by an extension of the pilot-tone system such that, if the stereo 'sum', stereo 'difference' and third-channel systems are given (after pre-emphasis) by Σ, Δ and T respectively, the composite signal is

$$\Sigma+\Delta \sin 2\omega t+T \cos 2\omega t+0.1 \sin \omega t$$
$$(\omega=2\pi\times 19 \text{ kHz}),$$

with the last term (the pilot-tone) representing 9% of the allowable deviation. If $T=0$ then this expression represents normal stereo.

The job of the demultiplexer (or decoder) in the receiver is to recover the audio channels from this signal. This is done almost universally by synchronous detection, in which the signal is

multiplied by a 38 kHz waveform, itself derived from the received pilot-tone (see Fig. 1). In principle, three-channel reception requires only the addition of another multiplier (or demodulator) to a stereo decoder. In practice, however, some of the shortcomings of existing decoder designs become more apparent in three-channel reception, and an improvement in performance is needed.

Common Shortcomings in Stereo Decoders

Two main areas of weakness can be identified. One of these lies in the demodulators. As it is simpler to do so, they generally use a switching circuit which has the effect of multiplication by a square-wave rather than by a true sinewave. This means that audible outputs result not only from the modulated 38 kHz sub-carrier but also from any signals which may be present near odd harmonics of 38 kHz. Such signals include beats due to adjacent r.f. transmissions at a spacing of 100 kHz or 200 kHz, which give rise to 'birdies' (i.e. twittering noises). For the same reason there is also some increase in random noise level, but the susceptibility to adjacent-channel interference (ACI) is usually more important since it causes audible beats which ideally need not be heard at all. This affects stereo and three-channel reception alike. The usual solution offered (e.g. ref. 7) is to pass the multiplex signal through a 53 kHz low-pass filter, but (apart from the expense) such a filter is particularly difficult to design for three-channel reception, because any phase delay variations in the pass-band will cause crosstalk between Δ and T.

The other weakness concerns the phase accuracy and stability of the 38 kHz waveform supplied to the demodulators. In stereo transmission and reception its phase is not very critical, since the effect of a phase error ϕ is only to change the gain of the signal by $(1 - \cos\phi) \sim \frac{1}{2}\phi^2$, which is usually small. So no great effort is made to control it, and in fact phase errors of 10° are commonly found (causing 1.5% loss in Δ gain, or -42 dB crosstalk between 'left' and 'right' signals, which may be corrected elsewhere). But in three-channel working the phase must be more accurate, since there will be crosstalk between Δ and T of magnitude $(\sin\phi) \sim \phi$, which is larger (e.g. a 10° error causes -15 dB crosstalk between Δ and T). Besides this static phase error, the recovered sub-carrier is also subject to signal-dependent phase fluctuations, notably when strong high audio frequencies are transmitted. Nominally, the transmitted pilot-tone is

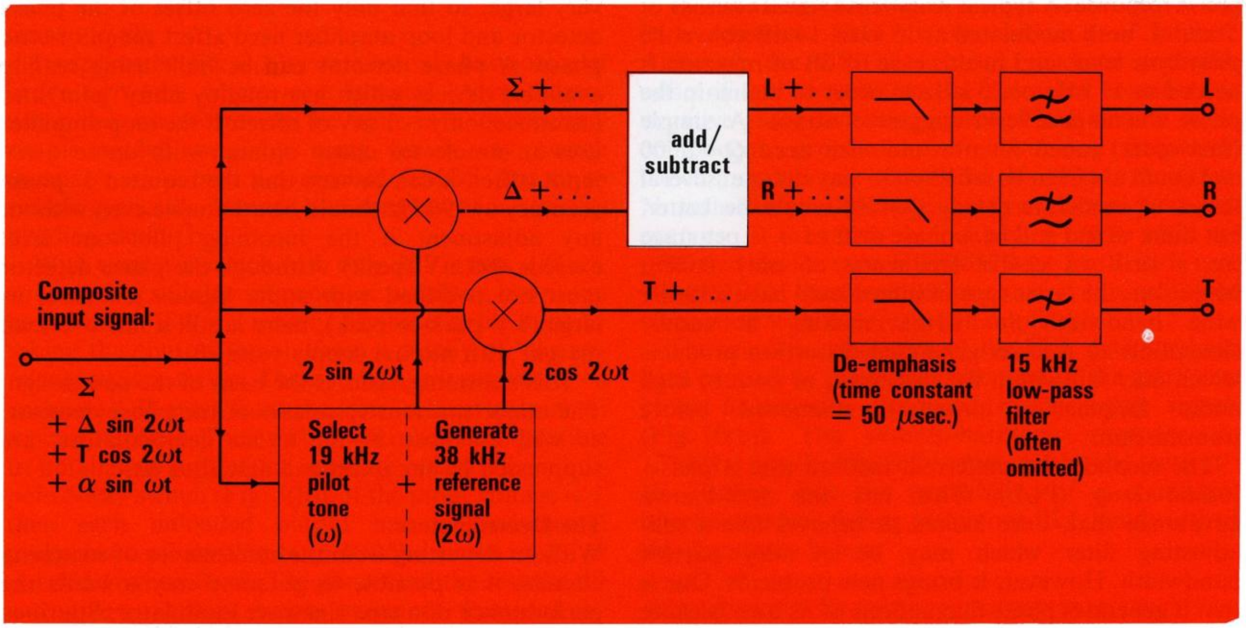


Fig. 1. The decoder or demultiplexer in a two-channel stereo or three-channel 'surround-sound' receiver is used to recover the original audio channels from the pilot-tone multiplexed transmission. Almost invariably the demultiplexer uses synchronous demodulation with the incoming signal multiplied by a 38 kHz signal derived from the 19 kHz pilot tone. In principle three- channel reception requires only the addition of a second synchronous demodulator to a stereo decoder. In practice, additional refinements are needed to overcome the shortcomings of many existing decoders.

protected by a 4 kHz 'gap' in the spectrum on each side of 19 kHz, but, even so, frequencies beyond this 'gap' can never be perfectly rejected, and remain to cause phase modulation of the pilot-tone (and hence of the recovered sub-carrier) at the difference frequency. This in turn is heard as distorted crosstalk. High levels of phase modulation are often allowed to occur in stereo decoders (e.g. fluctuations of $\pm 15°$, measured at 38 kHz, have been found with a worst-case signal), but this cannot be accepted in three-channel operation.

As a rough guide, it is proposed that for good domestic three-channel reception quality the static phase error of the recovered sub-carrier should not exceed 2° (giving -29 dB crosstalk between Δ and T), while any signal-dependent phase fluctuations should have a peak value at least 10 dB lower—say ± 0.01 radian (measured at 38 kHz).

Methods for Separating the Pilot-tone

There are two established methods. The traditional one uses a bandpass filter, followed by a frequency doubler to obtain 38 kHz. Unfortunately, this is scarcely adequate for three-channel reception, since a filter selective enough to give the required rejection of audio signals will not then be stable enough in its phase response. A typical worst-case signal consists of Σ and Δ, both modulated at 15 kHz, 14 dB above the pilot-tone level, and in this case 61 dB of rejection is needed at 15 kHz or 23 kHz in order to maintain the phase fluctuation level suggested above. A simple (first-order) tuned filter would then need $Q \sim 2{,}500$ and could not even be relied on to stay in tune; several stages of moderate Q (say $Q \sim 20$) would be better, but there would still be a phase drift of $\sim 1°$ per stage for a drift in centre frequency of only 0.04%. Moreover, the latter type of filter would have a rather wide bandwidth (~ 1 kHz), making it unduly susceptible to receiver-generated distortion products which may fall within the passband, or even to high audio frequencies imperfectly removed before transmission.

The second, and preferred, method uses a phase-locked loop (PLL). This has the well-known advantage that, once locked, it behaves like a self-adjusting filter which may be of very narrow bandwidth. However, it brings new problems. One is that it generates phase fluctuations of its own because of phase noise in the voltage-controlled oscillator (VCO); these increase as the bandwidth is reduced. Another is that, like the audio demodulators, the phase detector may have spurious responses. For instance, the usual squarewave-switching detector is only 9.5 dB less sensitive at 57 kHz than at 19 kHz, and so the loop could be appreciably disturbed if this frequency came to be used for transmitting signalling tones. In addition various forms of imbalance cause weaker responses at d.c. and near 38 kHz, so that low-frequency signals in Σ, Δ or T to some extent cause phase fluctuations directly.

Most stereo decoder designs, including the standard integrated circuits, now use the PLL method. Nevertheless, they rarely achieve its full potential. Very narrow loop bandwidths cannot be used because the VCO must then be very carefully preset if its frequency is to lie within lock-in range. High d.c. loop gains cannot be used either, lest in the quiescent state the frequency drifts out of lock-in range. Typically, therefore, one finds rather large loop bandwidths used (~ 300 Hz), giving only some 40 dB rejection at 15 kHz or 23 kHz, while in addition the recovered sub-carrier phase is sensitive to the VCO frequency adjustment.

These problems can be solved if the PLL works in two modes, one for acquiring lock and one for normal running. This proves to be quite easy to do. In the former ('fast') mode the bandwidth can be wide, while in the latter ('slow') mode the d.c. loop gain can be very large, so that only the zero offset of the phase detector and loop amplifier need affect the pilot-tone phase. A phase detector can be built using readily available devices which has roughly unity 'gain' and introduces only ~ 1 mV of offset; if the loop amplifier uses an unselected cheap op-amp with up to 6 mV input offset, it can be seen that the required 1° phase accuracy at 19 kHz should be attainable even without any adjustment if the incoming pilot-tone level exceeds 400 mV (peak). Although the phase detector must not overload with audio signals ten times as large (8 V peak-to-peak), there is still a fair overload margin with normal supply voltages.

This reasoning, then, is the basis of the new design. The major spurious responses of the phase detector, as well as those of the audio demodulators, are suppressed by the technique described here.

The Demodulators

Without departing from the convenience of switching circuits, it is possible to go some way towards the performance of a true sinewave modulator. Spurious responses will not occur if the corresponding harmonics are not present in the effective modulating waveform. Such a waveform is that of Fig. 2(a), obtained by sampling-and-holding a sinewave N

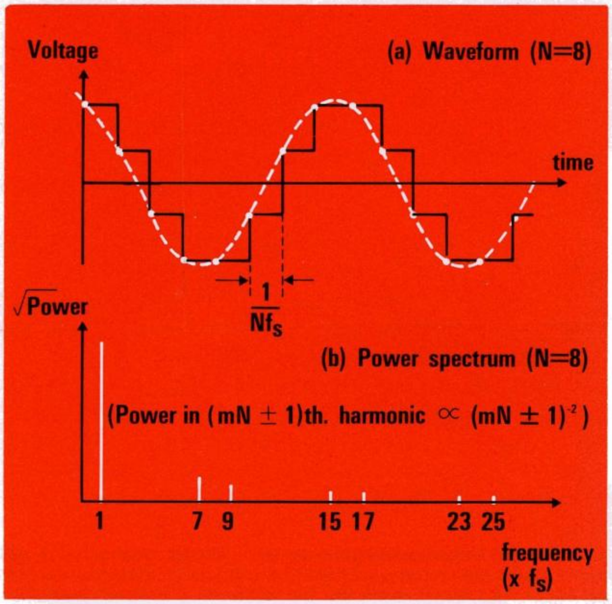


Fig. 2. To combine the convenience of switching circuits with the performance of a true sinewave modulator it is necessary to remove the harmonics that are caused by switching the modulating waveform. (a) Shows how a near sinewave is reconstructed by the use of a sampling-and-holding technique, repeated N times a cycle. (b) Output spectrum of such a waveform contains only odd-order harmonics of $mN \pm 1$ form where m is an integral number.

times a cycle.

Its spectrum (Fig. 2(b)) contains only harmonics of order $(mN \pm 1)$, where m is integral. A modulator using this waveform can be built in a particularly simple-minded way by using an N-way switch (available as a binary-controlled CMOS device) to switch cyclically between the outputs of a resistive dividing chain (Fig. 3).

With even N a switch with only $N/2$ inputs suffices, so long as a suitable sequence of switching codes can be provided. If N is a power of 2 this is easily done by taking the control signals from a family of 'cosine'-related square-waves. These in turn are generated via exclusive-OR gates from the 'sine'-related outputs of a binary divider. Thus with $N=8$ it has been possible to obtain two outputs of opposite polarity from a 2-pole 4-way switch (Fig. 3(b)). If the input voltage is V then each unloaded output has a fundamental component of amplitude $\pm V \times N/\pi \times \tan \pi/N$, or $\pm 1.055\ V$ when $N=8$. The purpose of the resistor R2 is to give a constant output resistance. Provided the outputs are not too heavily loaded the 'gain' is accurately defined and the distortion is low.

Circuit designs have appeared elsewhere, apparently independently, which contain modulators based in effect on the case $N=6$: for example, the Motorola TCA4500A stereo decoder IC[8], and new stereo encoder designs from Harris and Radiometer. In these the approach seems to have been purely one of cancelling the third-harmonic response, and does not imply awareness of the case with general N. In fact a value of $N=6$ would not be the best starting-point for three-channel decoding, because additional switching transitions have to be interpolated for the second (quadrature) demodulator.

Switching methods other than CMOS might be more suitable for higher switching rates or for large-scale integration, notably the current-switching technique using long-tailed pairs.

The actual measured response of the new ($N=8$) demodulators in the final circuit (see Fig. 9(b)) shows that unwanted harmonics below the seventh are present at roughly -50 dB level, an improvement of some 40 dB over conventional square-wave switching (Fig. 9(a)). In use it can be assumed that ACI more than 250 kHz off-frequency is already well attenuated in the receiver's IF section, provided that the receiver is correctly tuned, so that the seventh and higher harmonic responses are of no consequence. Thus the need to precede the demultiplexer with a low-pass filter is substantially eliminated.*

The same circuit (Fig. 3(b)) has been used in the PLL detectors. Here its zero offset is of interest. Although the CMOS switches generate negligible offset if left to settle, in use there are considerable negative-going switching spikes. These have a mean component, proportional to switching frequency, which amounts to -30 mV in the 19 kHz detectors. Nonetheless, the d.c. match between the two push-pull outputs is found to be within ± 1 mV with the preferred device type HEF4052B (± 2 mV with type CD4052BE). Excellent performance is thus obtained when used with a differential amplifier.

The spurious responses of the PLL detector have been plotted by direct measurement of the pilot-tone phase fluctuations produced by incoming signals (Fig. 10(b)). The seventh-harmonic response (at 133 kHz) may not be fully protected by the receiver's IF filtering, but the probability of a significant interference beat falling within one PLL bandwidth of this frequency is low.

* This may not be strictly true where there is interference so strong as to deeply modulate the wanted signal, since the discriminator output may then contain components outside the half-bandwidth of the IF filters. A relatively simple filter, rejecting 250 kHz upwards, might prove useful in this extreme case.

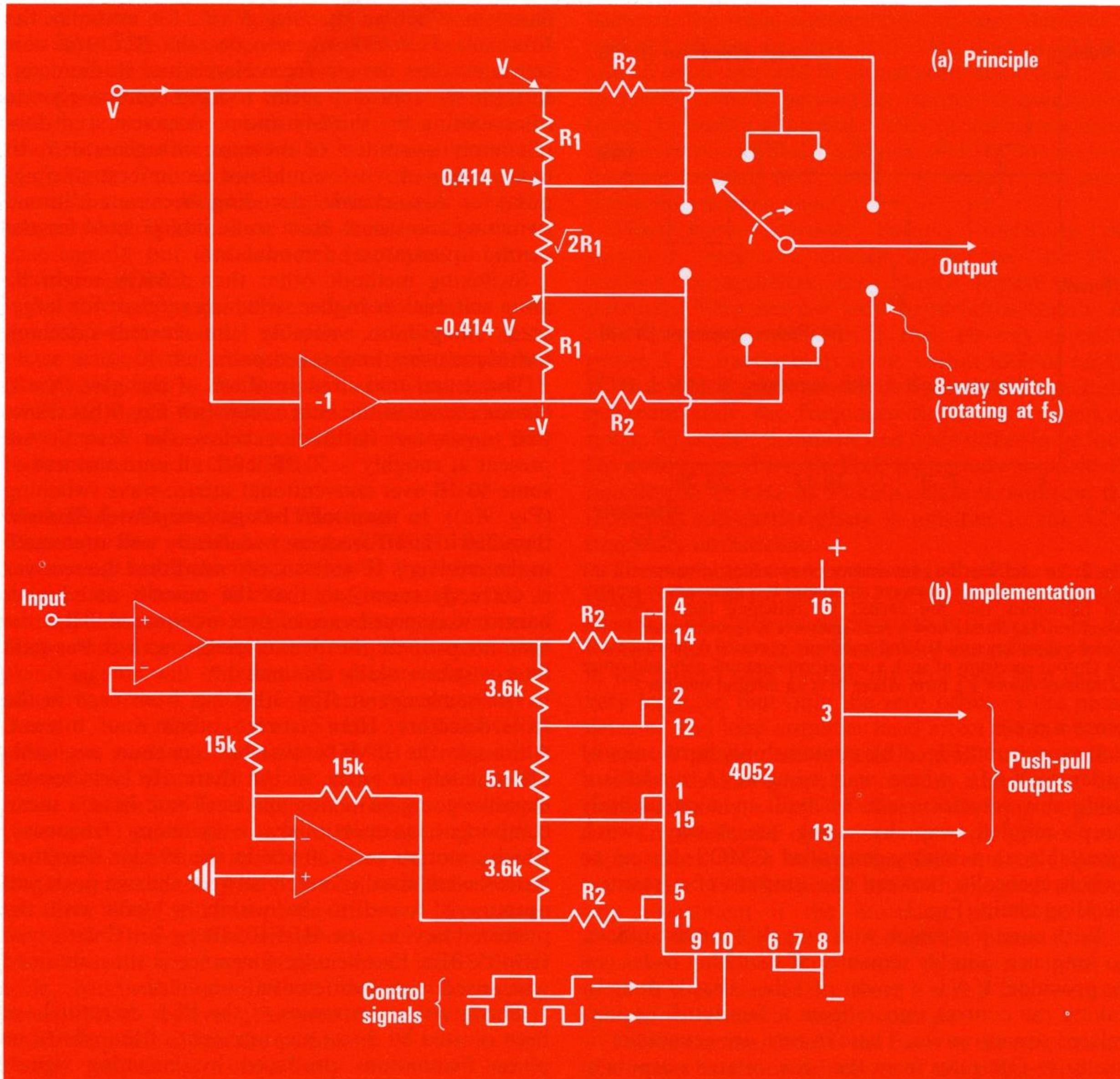


Fig. 3. (a) Principle of a simple switching modulator using a waveform reconstructed as in Fig. 2, based on an *N*-way switch available in integrated cmos form switching cyclically between the outputs of a resistive dividing chain. (b) Method of implementing such a modulator. With $N=8$ it is possible to obtain two outputs of opposite polarity from a two-pole, four-way electronic switch.

The Phase-locked Loop

The sub-carrier recovery system is shown in Fig. 4. The phase detector circuit is that already described; its 'gain' (maximum d.c. out ÷ peak a.c. in) is 1.055 (unloaded) when measured across both outputs. The VCO uses an emitter-coupled oscillator circuit which has good temperature stability and much lower phase noise than the more familiar multivibrator used, for example, in ref. 9. Its sensitivity is 1% frequency change per volt at the loop amplifier output. There is a preset frequency adjustment, but it is necessary only to set it within loop capture range (±4%).

Most simple PLL's have a second-order response. This includes stereo decoder ICs like the TCA4500A.

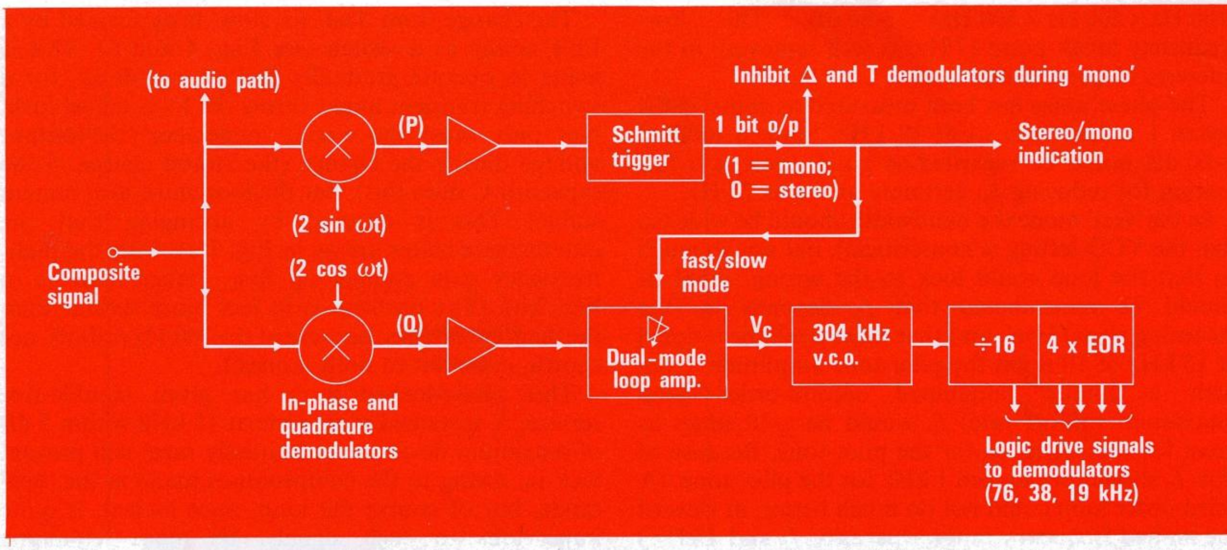


Fig. 4. Sub-carrier recovery system using phase-locked loop. The emitter-coupled voltage-controlled oscillator has been designed for good temperature stability and much lower phase-noise than where a more conventional multivibrator arrangement is used. A preset frequency adjustment control is provided but this needs be set only to within the 4% loop capture range.

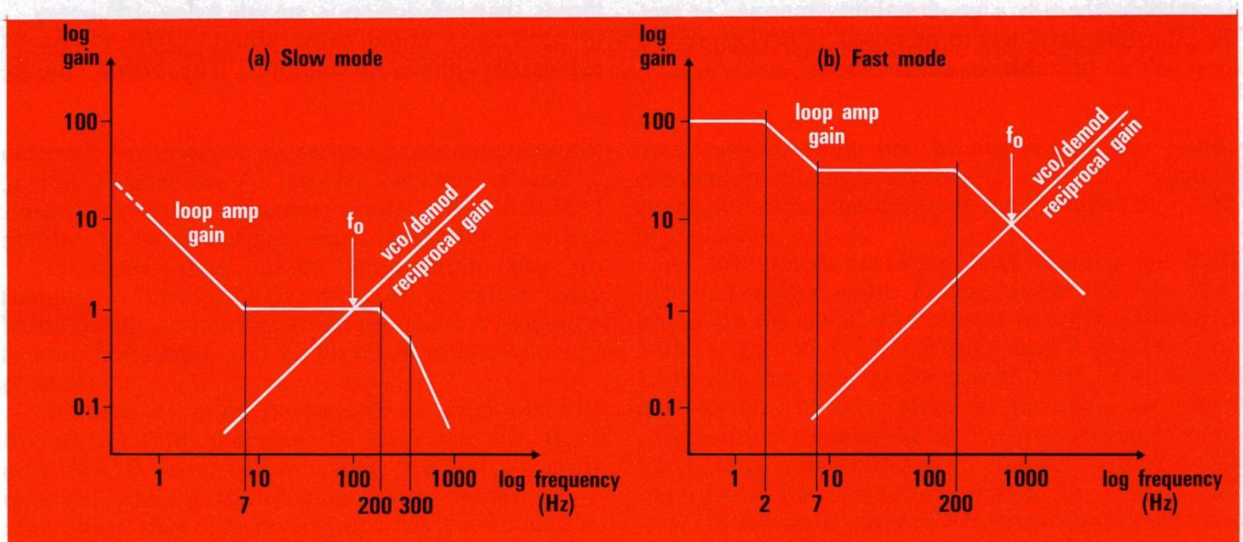


Fig. 5. Slow and fast mode responses of the phase-locked loop. In the fast mode the loop bandwidth needs to be reasonably wide so that the adjustment of the voltage-controlled oscillator is non-critical, but not so wide that the loop locks on to the wrong signal. A loop bandwidth of 900 Hz provides a suitable compromise and, in practice, the pilot-tone level has to be raised some 17 dB before the loop becomes unstable in either mode.

To get 61 dB rejection for beat frequencies of 4 kHz (as suggested earlier) the loop bandwidth must then be about 100 Hz or less. However, there is plenty of room between 100 Hz and 4 kHz to place an additional roll-off point, giving a third-order loop with comfortably superior beat rejection combined with good damping. (A high Q is to be avoided because it imparts a definite timbre to the sound of any crosstalk effects associated with phase fluctuations.) This has been done in the normal-running mode (Fig. 5(a)); the loop bandwidth (f_0) is 100 Hz, and at 4 kHz the rejection is $(4\,\text{kHz})^3 \div$

(100 Hz × 200 Hz × 300 Hz) $\simeq$80 dB. The low-frequency break point (7 Hz) is well removed, in the interests of good damping.

The phase noise has been measured as some 75 dB below 1 radian (r.m.s.) at 19 kHz. Since a level of −66 dB would be regarded as good, there is a fair margin for reducing f_0, certainly as far as 25 Hz.

In the 'fast' mode the bandwidth should be wide (so that the VCO setting is non-critical), but not so much so that the loop could lock to the wrong signal. It should be remembered that f_0 depends on the incoming signal level, and that there could be a signal at 15 kHz of 10 times the pilot-tone amplitude. Even with a totally undamped second-order loop (maximum roll-off rate) f_0 would be $\sqrt{10}$ times as great for this signal as for the pilot-tone. Because of this, f_0 should not exceed 1 kHz for the pilot-tone. (A third-order loop could not do much better, in view of the limited frequency range separating f_0 and 4 kHz.) So some compromise is involved, and f_0 was set to 900 Hz (Fig. 5(b)). Because this frequency falls on the slope of the 200 Hz roll-off the Q is fairly high, but nonetheless there is a good stability margin: in fact the pilot-tone level has to be raised 17 dB before the loop will go unstable, in either mode.

The change from 'fast' to 'slow' is performed by a FET acting as a switch (see Figs. 4 and 6). This is made to operate gradually and after a delay, for a switching transient might throw the loop out of lock. Not only must the circuit 'remember' the output voltage during the change (the stored charge in the capacitor C does this), but the loop must also remain stable! This is ensured by arranging that the intermediate states are as in Fig. 7(a), with the high-frequency gain coming up first, rather than as in Fig. 7(b). (The problem is in fact more serious than the diagram implies, because of the 200 Hz roll-off, not shown, common to both modes.)

This phase-locked loop has given trouble-free service. A sustained monotone at 15 kHz within 5 dB of maximum deviation (admittedly rare) will prevent lock-in, owing to the compromises made in the 'fast' mode, but of course the loop, once locked, is quite impervious to such signals. Its main remaining weakness is due to low-frequency audio breakthrough. The spurious responses at 0 kHz and 38 kHz are some 52 dB down (see Fig. 10(b)), which means that full-deviation signals below frequency f_0 in Σ, Δ or T could cause about ± 0.05 radian of sub-carrier phase modulation. The effect can be

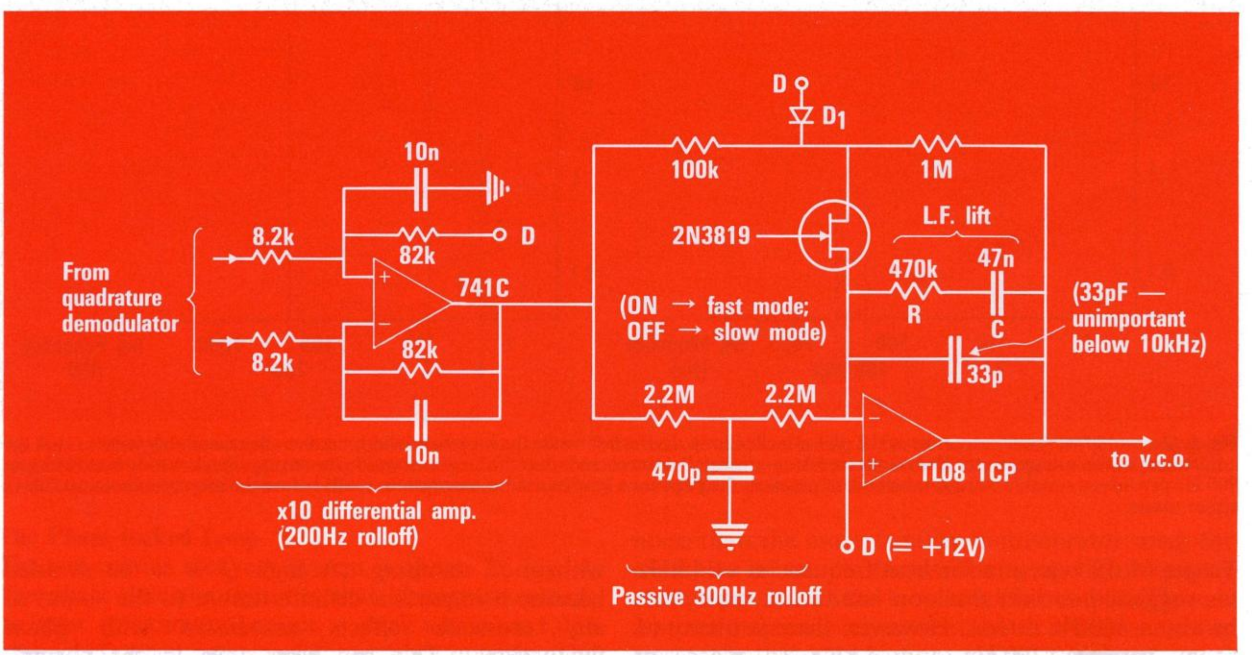


Fig. 6. The change from fast to slow modes of the phase-locked loop is brought about by the switching action of a field-effect transistor. The switch is made to act gradually and only after a delay to prevent switching transients from throwing the loop out of lock. The circuit has to 'remember' the output voltage during the change, while at the same time the loop must remain in a stable condition.

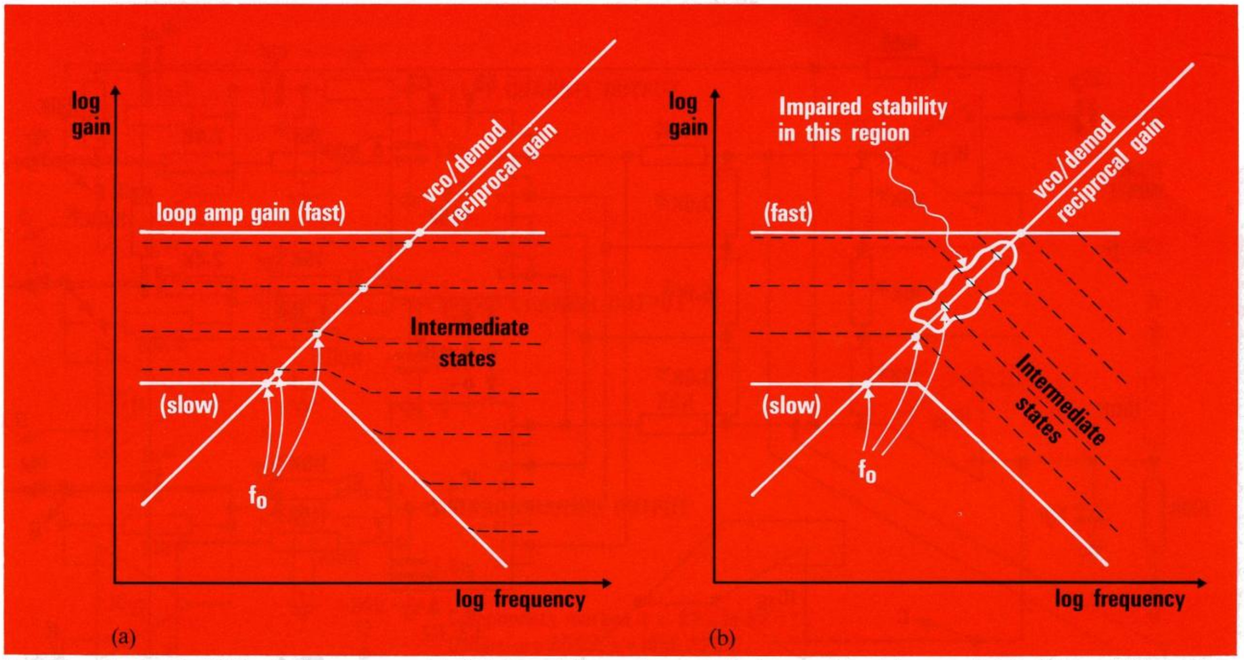


Fig. 7. In order that the FET switching arrangement of Fig 6 meets its requirements, the intermediate states are as in (a) with the additional high-frequency gain applied first, rather than as indicated in (b). The problem is rather more serious than implied in the diagram because of the 200 kHz roll-off which is not depicted. However the phase-locked loop has, in practice, given trouble-free service.

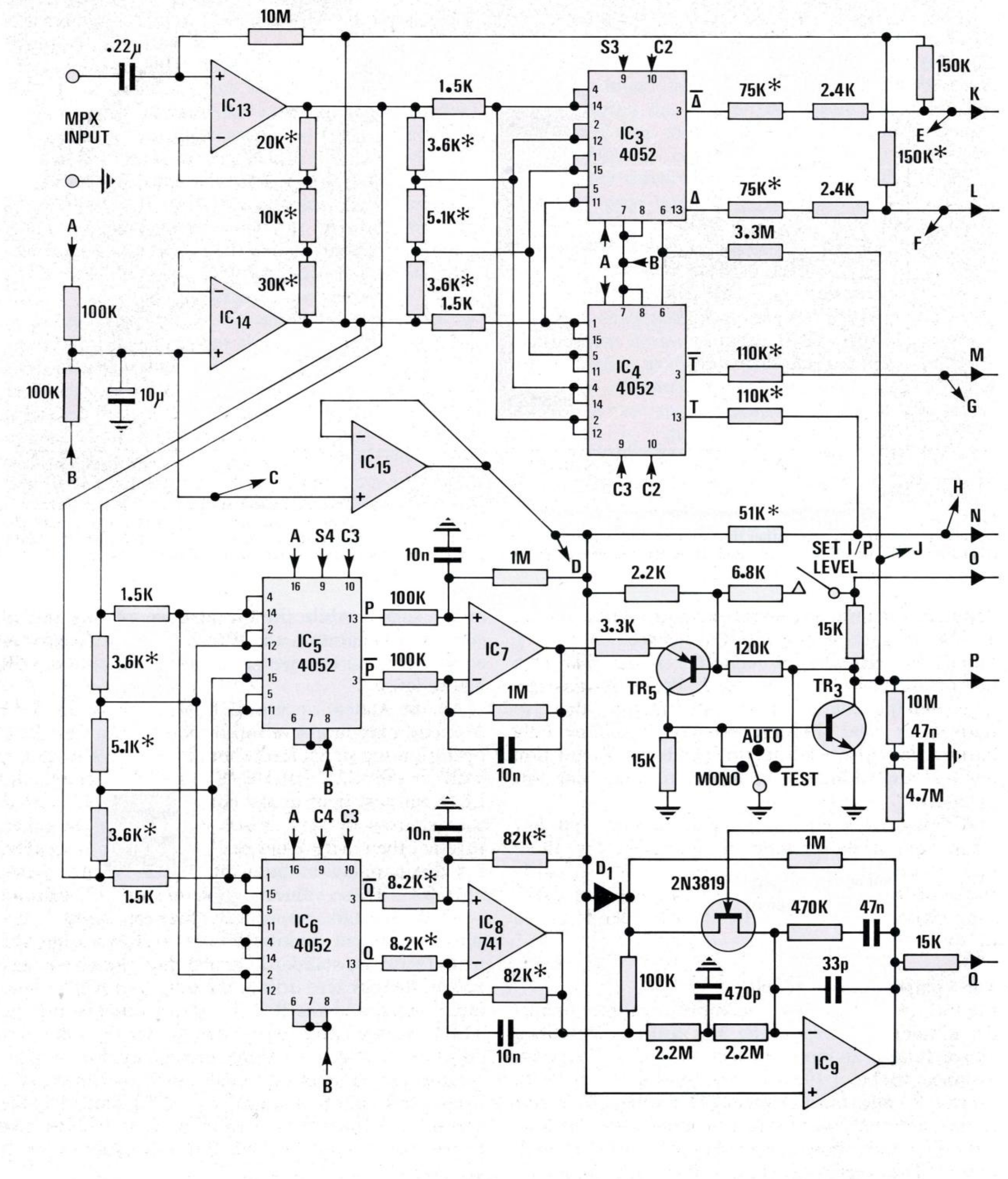
detected, for instance, as second-harmonic crosstalk at low frequencies (<100 Hz) between Δ and T signals. Nevertheless its magnitude is less than in a number of other designs tested. Probably it is less objectionable than phase modulation due to frequencies near 19 kHz, because it causes only harmonically-related distortion products. A reduction in loop bandwidth will be useful in minimising this problem.

Although it would be possible to filter out low frequencies from reaching the loop detector, those near 38 kHz (equally important) are not so easily removed without introducing poorly-defined pilot-tone phase shifts, so no attempt has been made to do so.

The Complete Basic Circuit

Figure 8 shows a complete three-channel demultiplexer incorporating the features described above. It takes an input in which the pilot-tone level is assumed to be ± 100 mV, and produces 'left' (L), 'right' (R) and third-channel (T) outputs. For best results, especially when tape-recording, a 15 kHz low-pass filter (not shown here) should be used in each output. The overall gain (to L or R) is unity for a pure mono signal, while (in the absence of any settled convention relating L and R to Σ and Δ) a T signal of equal deviation emerges from the T output at a 3 dB higher level.

At the stated operating level there is an 8 dB overload margin above full deviation. For best PLL operation the signal level should be correct to within 3 dB; if the 'SET I/P LEVEL' switch is closed the LED will just light at the correct level. If all worst-case sources of zero offset in the PLL are taken together then there could be 2° phase error at 19 kHz, but a square-law addition of 'typical' figures yields only 0.4°. These values correspond to -23 dB and -37 dB crosstalk, respectively, between Δ and T. If a zero adjustment is introduced for IC8 (using the manufacturer's standard circuit) then the phase may be adjusted for zero drift at the output of IC9 with no input signal and the PLL in 'slow' mode (using the TEST switch position shown). Better still, this adjustment might be done merely by listening to suitable transmitted test signals. Such a setting results in crosstalk figures of typically -40 dB, limited by the spread in propagation delays in ICs 3–6. In this respect the IC type HEF4052B is once again found to give the best results.


10M
.22μ
MPX
INPUT
IC13
20K*
10K*
30K*
IC14
A
100K
100K
10μ
B
1.5K
3.6K*
5.1K*
3.6K*
1.5K
S3
C2
IC3
4052
75K*
2.4K
150K
K
E
150K*
75K*
2.4K
L
F
3.3M
A
B
IC4
4052
110K*
M
G
110K*
C3
C2
IC15
C
H
51K*
N
SET I/P
LEVEL
J
O
A
S4
C3
10n
1M
D
2.2K
6.8K
1.5K
IC5
4052
P
100K
100K
IC7
3.3K
120K
15K
P
3.6K*
1M
TR5
TR3
10M
47n
5.1K*
10n
B
15K
AUTO
MONO
TEST
4.7M
A
C4
C3
10n
82K*
3.6K*
1M
1.5K
IC6
4052
8.2K*
8.2K*
IC8
741
D1
2N3819
470K
47n
15K
Q
82K*
100K
470p
33p
10n
2.2M
2.2M
IC9
B

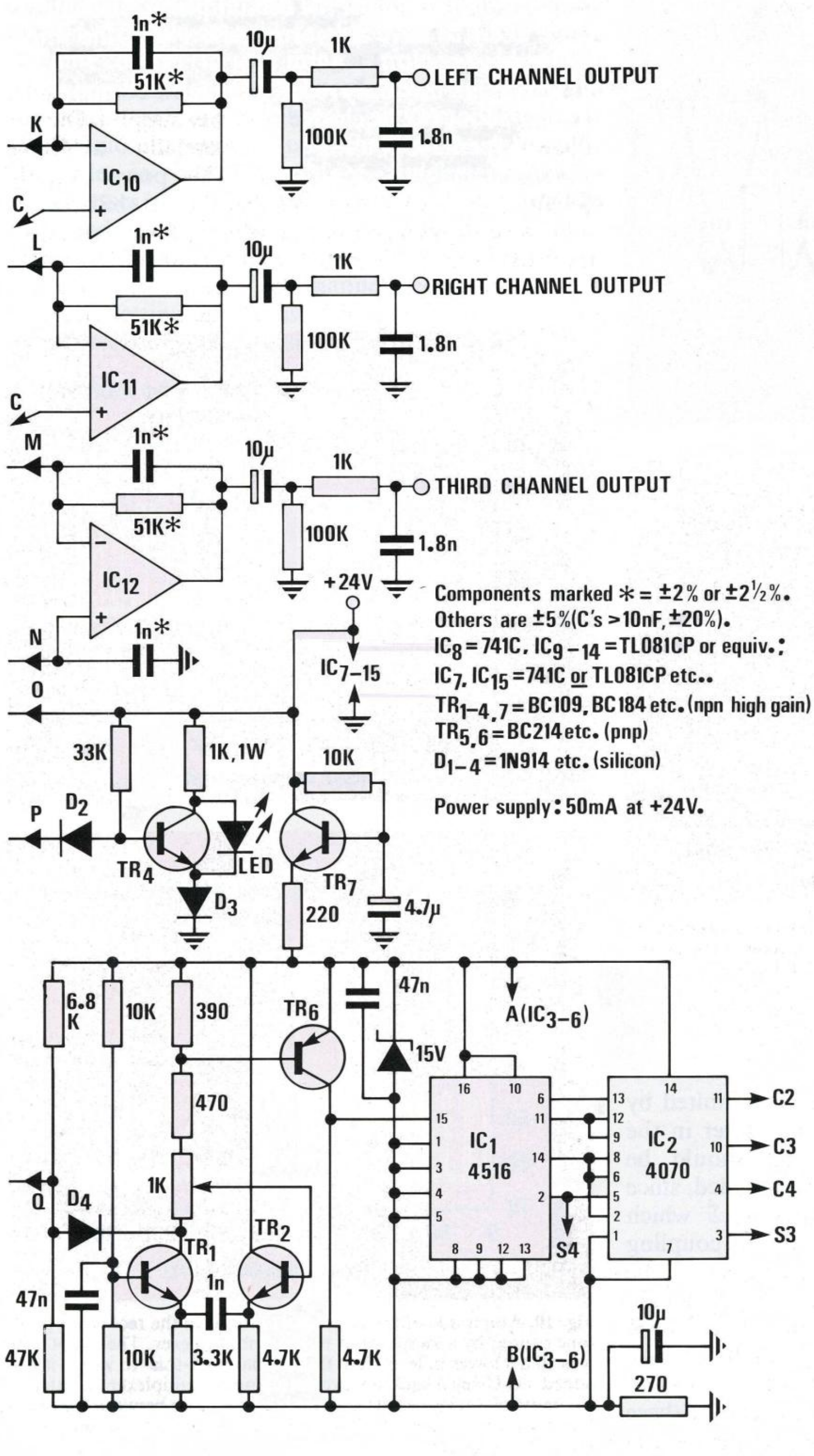


Fig. 8. Basic form of the complete three-channel multiplexer incorporating all the features outlined in this Section. With a pilot-tone input level of about ± 100 mV it produces 'left' (L), 'right' (R) and third-channel (T) outputs. For best results when tape recording a 15 kHz filter should be connected in each output. With such a demultiplexer performance is essentially limited only by the basic FM tuner. Additional facilities could be added.

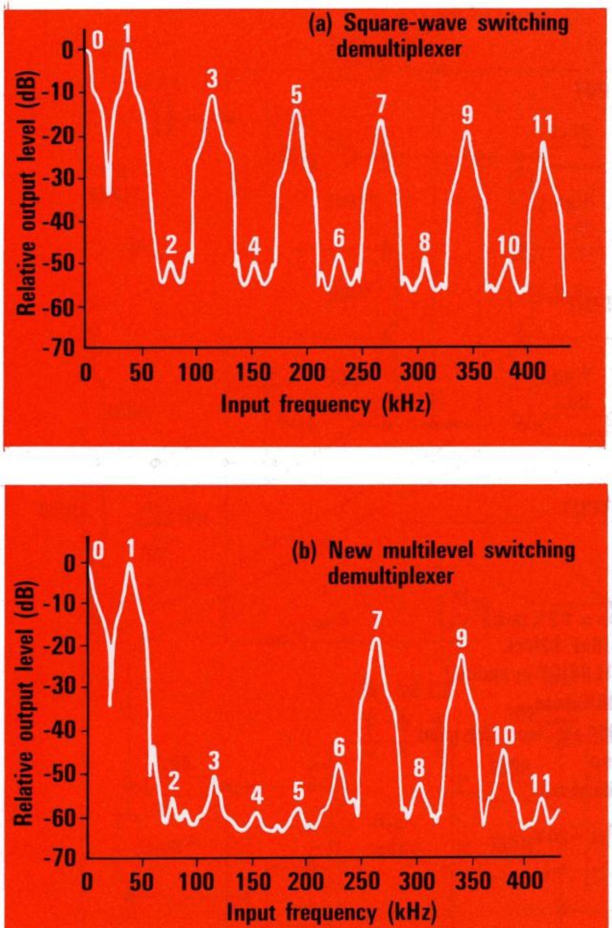


Fig. 9. Response of demultiplexers ('L' output) when a swept input signal is added to the pilot-tone. The total amplitude of the output signal was measured via a 15 kHz low-pass filter which, together with the de-emphasis, is responsible for the characteristic shape of each peak. (a) Response of square-wave switching demultiplex. (b) Response of the new multilevel switching demultiplexer.

Fundamentally the performance is now limited by the FM tuner. Any 53 kHz 'anti-birdie' filter in the multiplex output is not needed and should be bypassed. AC couplings should also be avoided, since they cause low-frequency phase shifts in Σ which reduce the L–R separation. (The input coupling shown in Fig. 8 allows 46 dB separation to be maintained down to 15 Hz.) Any high-frequency gain droop in the tuner's output will affect L–R separation, while phase non-linearity results in Δ–T crosstalk; typically -35 dB crosstalk is to be expected in either case. It is better to correct for these faults by inserting a filter with an upward 'step' in the response than to adjust the demultiplexer, since this is more likely to have the correct effect over the whole frequency band.

The design uses only readily-available components, costing about £12 (exclusive of power supply). Output filters (perhaps using commercially available packages) might add £3–£4. The power supply (50 mA at +24 V) should be d.c. stabilized, as otherwise slow mains voltage fluctuations can affect the PLL.

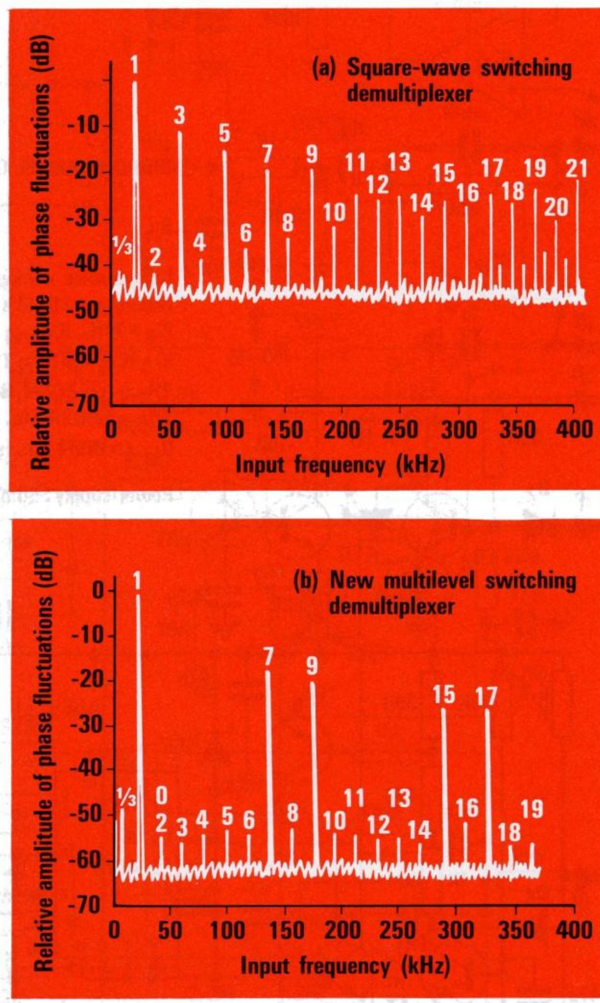


Fig. 10. Amplitude of the phase fluctuations in the recovered pilot-tone caused by a swept input to the demultiplexer. The swept tone was 10 dB lower in level than the normal pilot-tone to which it was added. (a) Using a square-wave switching demultiplexer (as fitted to a commercial receiver). The random phase noise between the peaks is exaggerated by the quasi-peak detector. (b) Using the new multi-level switching demultiplexer.

Only the basic form of the circuit is shown here. It is not difficult to add facilities such as an input volume control (to save using a triple-ganged control elsewhere), pilot-tone cancellation (to ease the demands on output filters), or demodulation of additional multiplexed audio channels (as defined for example in refs. 3–5). More fundamentally the weaknesses of the sub-carrier separation circuitry have been indicated. The aim of this work, however, was merely to produce a reasonably simple design capable of good domestic three-channel reception without critical setting-up adjustments; it is incidentally capable of excellent stereo reception as well.

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A biographical note and photograph of Dr Windram appear on page 41.

JONATHAN HALLIDAY, BA, Ph.D.
A biographical note and photograph of Dr Halliday appear on page 60.

APPENDIX

Adaptive Aerial Arrays—A Theoretical Introduction

by M. D. Windram and J. Halliday

Synopsis
An earlier Section of this *IBA Technical Review* describes the practical design and implementation of adaptive aerial arrays which are able to adjust themselves automatically to meet particular constraints. Such constraints include optimum rejection of unwanted signals causing co-channel interference (CCI). This Appendix presents a rigorous theoretical introduction to adaptive aerials when based on a linear, uniformly-spaced array of identical elements, together with possible adaptation algorithms. It is shown that significant limitations exist on adaptation rates caused by correlations and coherence between wanted and unwanted signals (whether in the form of CCI or deliberate electronic jamming). For this reason the theory and design of the measuring system is an important part of adaptive-array design for a wide variety of applications.

A requirement in broadcasting, communications and radar is an aerial array that has a response pattern that includes a number of deep (rejection) nulls, or nulls which will track interference or jamming. The limitations on the use of conventional aerial arrays have been set out in an earlier Section.

Adaptive arrays consist in general of a number of aerial elements, not necessarily identical, coupled via amplitude-adjusting and phase-shifting networks and a power combiner to provide an integrated signal output.

The general case for such an array is complicated. In this Appendix the theory is restricted to the simple case of the uniformly-spaced linear array of identical aerial elements. This case covers the approach adapted for the IBA SABRE array and is typical of arrays used in practice for other applications.

Theory of Linear Uniformly Spaced Adaptive Array

The type of array being considered here is shown in Fig. 1. The effect of noise and mutual coupling at the antenna elements has been omitted from the earlier parts of the discussion to simplify the analysis.

If we define $W_n = X_n + jY_n$ = complex weighting coefficient for the nth element, V_{kn} = the voltage at the nth element for a source at angle θ_k and T_k = the combined complex output for a source at angle θ_k, then

$$T_k = \sum_{n=1}^{N} V_{kn} \qquad (1)$$

and

$$V_{kn} = V_k W_n \exp\left(2\pi \frac{jnd}{\lambda} \sin\theta_k\right) \qquad (2)$$

so that

$$T_k = \sum_{n=1}^{N} V_k W_n \exp\left(2\pi \frac{jnd}{\lambda} \sin\theta_k\right) \qquad (3)$$

If we define

$$\psi_k = \frac{2\pi d}{\lambda} \sin\theta_k \qquad (4)$$

then

$$T_k = V_k \sum_{n=1}^{N} W_n \exp(nj\psi_k) \qquad (5)$$

If θ_1 is the bearing of the wanted signal and the θ_k ($k \neq 1$) are the bearings of the unwanted signals, then we can see that, by suitable choice of the values of W_n, $T_k = 0$ for the unwanted signals. It is the coefficients W_n which the adaptive array varies to null the interference.

Note that $\psi = \psi(\theta_k, \lambda)$ in eqn. 4 above for a given

array, so that sources at different frequencies on the same bearing can be considered to be on the same frequency but at different bearings. This property is important in an array which is required to operate on several frequencies simultaneously.

It is useful at this point to introduce vector and matrix notation for the expressions derived in eqns. 1–5 above:

T is a column vector with elements $T_1 \ldots T_K$

M is a matrix with elements M_{ab}

T* and **M*** are the appropriate complex conjugates

$\tilde{\mathbf{T}}$ represents the row vector form of **T**

$\tilde{\mathbf{M}}$ is the transpose of **M** with elements M_{ba}

The expression for **T** can therefore be written in matrix form as

$$\begin{bmatrix} T_1 \\ | \\ | \\ | \\ T_K \end{bmatrix} = \begin{bmatrix} V_1\, e^{j\psi_1} & --- & V_1\, e^{jN\psi_1} \\ | & & | \\ | & & | \\ | & & | \\ V_K\, e^{j\psi_K} & --- & V_K\, e^{jN\psi_K} \end{bmatrix} \begin{bmatrix} W_1 \\ | \\ | \\ | \\ W_N \end{bmatrix} \qquad (6)$$

T_1 = wanted and $T_2 --- T_K$ = unwanted. If we further define

$$U_{kn} = V_k\, e^{jn\psi_k} \qquad (7)$$

then

$$\mathbf{T} = \mathbf{UW} \qquad (8)$$

where **U** is an $N \times K$ element complex matrix with K = number of sources including the wanted source.

The aim with an adaptive array is to fix or optimise the wanted signal while minimising the level of interference or jamming. This corresponds to reducing all the elements of the vector **T** to zero except T_1, the wanted signal.

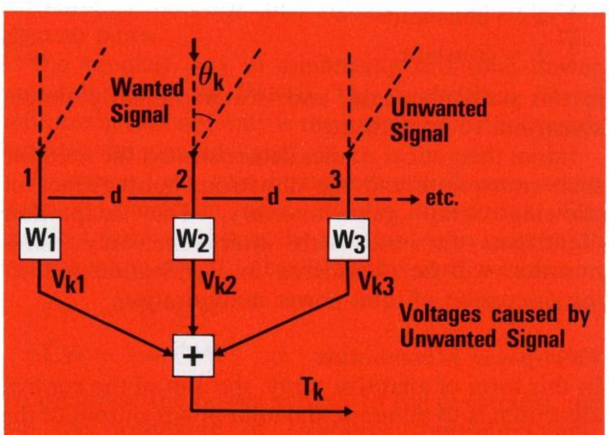

Fig. 1. Principle of operation of a phased aerial array, where N is the number of elements, d is the spacing between the elements and λ is the wavelength.

Given the matrix relationship of eqn. 8, there are three possible solutions:

(*a*) We can see that if $N = K$, i.e. the number of sources of interference or jammers is one less than the number of elements in the array, then

$$\mathbf{W} = \mathbf{U}^{-1}\,\mathbf{T} \qquad (9)$$

so that, for T_2 to $T_K = 0$, **W** is fully defined (other than the constant factor which is array gain). This means that, for this particular case, the values of the weights to be used can be determined uniquely (other than the constant factor) once the directions of the interfering sources are known.

(*b*) If $K > N$, then there are insufficient degrees of freedom and T_2 to T_K cannot all be reduced to zero simultaneously.

(*c*) If $K < N$, then again a unique inverse does not exist. An inverse could be calculated by adding extra 'fictitious' sources of interference to make $K = N$. Since the directions of these fictitious sources can be made entirely arbitrary, it follows that there is a whole range of matrices and therefore a whole range of weights which correspond to no interference.

Matrix inversion also has the major disadvantage that, being an open-loop system with no feedback from the output of the array, it takes no account of inaccuracies in the directions θ_k or in the array itself. This is therefore not in general a satisfactory method of determining the weights to use, even if the directions θ_k are known.

The technique does, however, have the one advantage of not requiring any measurements on the received signals and can therefore be used to compute the starting weight vector **W** for other optimisation processes.

We therefore require a general method of minimising $T_2 \ldots T_K$ while maintaining or optimising T_1. To remove the effects of source movement, array inaccuracy etc., the method must be closed loop. There are two distinct approaches to this, depending on whether there is any possible technique of distinguishing wanted signal from interference. The first approach, and that which is considered in some depth in this paper, is to minimise the interference only. The second approach, which is much more widely covered in the literature,[1] is to assume that there is no distinction between signal and interference, and therefore to minimise the total power.

Interference-only Minimisation

The situation studied here in some depth is that in which it is possible to recognise the difference between the wanted signal and the interference. This has several advantages:

(i) it is possible to recognise and therefore minimise the interference without affecting the wanted signal.

(ii) there is no natural tendency to minimise the wanted signal.

(iii) the system should be more rugged in its response to changes in the wanted signal.

It is useful here to introduce the concept of the error function

$$E = \sum_{k=2}^{K} |T_k|^2 \quad \ldots \quad (10)$$

It is likely in many communications applications that some form of AGC will be involved, so that the quantity really measured is

$$\epsilon = E/|T_1|^2 \quad \ldots \quad (11)$$

If we resolve the complex weights W_n into $2N$ real variables x_k, where

$$W_n = x_{2n-1} + jx_{2n} \quad \ldots \quad (12)$$

then, from eqns. 5, 10 and 11, we can see that both E and $|T_1|^2$ are quadratic functions of $(x_1 \ldots x_{2N})$.

The function $\epsilon(x_1 \ldots x_{2N})$ is therefore not quadratic so that it is not a satisfactory function to minimise as it stands. Since this is the function which is most readily measured and which we must minimise to maximise signal to interference ratio, it is useful to find a method of keeping $|T_1|^2$ constant so that ϵ is quadratic. This can be done by introducing the concept of a constraint. The constraint takes the generalised form of a vector $\mathbf{C}$ such that

$$\tilde{\mathbf{C}}\mathbf{W} = \text{constant} \quad \ldots \quad (13)$$

In particular, if we define the constraint vector according to the relationship

$$C_n = U_{1n} \quad \ldots \quad (14)$$

where U_{1n} is defined in eqn. 7, then

$$T_1 = \sum_{n=1}^{N} U_{1n} W_n = \sum_{n=1}^{N} C_n W_n = \tilde{\mathbf{C}}\mathbf{W} = \text{constant} \quad (15)$$

and

$$\delta T_1 = 0 = \tilde{\mathbf{C}}\delta\mathbf{W} \quad \ldots \quad (16)$$

It is useful at this point to introduce the concept of controls. There are as many controls as weights, but whereas weights are not permitted to be changed independently for constant gain operation, the controls can be. The weights and controls are related by the following:

$$\mathbf{W} = \mathbf{W}_0 + \mathbf{MA} \quad \ldots \quad (17)$$

so that

$$\delta\mathbf{W} = \mathbf{M}\delta\mathbf{A} \quad \ldots \quad (18)$$

where $\mathbf{A}$ is the control vector, with independent elements or controls A_i.

If we determine the matrix $\mathbf{M}$ according to the relations

$$M_{ni} = -1/N \exp\{(1-n)j\psi_1\} \qquad n \neq 1 \quad (19)$$

$$M_{nn} = 1 - 1/N$$

then each element δA_i of $\delta\mathbf{A}$ adjusts the weights W_n such as to keep T_1 constant. Furthermore, if the starting values of W_n are defined as

$$W_n = A \exp(-nj\psi_1) \quad \ldots \quad (20)$$

then the main lobe of the array points to the wanted signal.

For a wanted signal on boresight, $\mathbf{M}$ reduces to

$$\delta\mathbf{W} = \frac{1}{N}\begin{bmatrix} N-1, & -1, & -1 \\ -1, & N-1, & -1 \\ -1, & -1, & N-1 \end{bmatrix} \delta\mathbf{A} \quad (21)$$

with a starting condition W_i = constant.

This corresponds to

$$\sum X_i = \text{constant}$$
$$\sum Y_i = \text{constant} \quad \ldots \quad (22)$$

in this particular case, and is a particularly simple constraint to implement.

Initial theoretical studies demonstrated the need for such constraints and the disastrous consequence of allowing array gain to vary. The adaptation algorithms for use with interference-only minimisation will be considered in the Section following discussion of total-power minimisation.

Total-power Minimisation

In this form of adaptive array, the aim of the control algorithm is to minimise the total power output of the aerial system, subject to some form of constraint which attempts to keep the wanted signal from being nulled. This technique is necessary when there is no

means of distinguishing signal from interference. The total power output of the array is proportional to

$$P_{TOT} = \left| \sum_{k=1}^{K} T_k \right|^2 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (23)$$

If we define

$$y = \sum_{k=1}^{K} T_k \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (24)$$

then

$$y = \sum_{k=1}^{K} V_k \sum_{n=1}^{N} W_n \exp(nj\,\psi_k) \quad \ldots \quad \ldots \quad (25)$$

so that

$$y = \sum_{n=1}^{N} W_n \sum_{k=1}^{K} V_k \exp(nj\,\psi_k) = \sum_{n=1}^{N} W_n S_n \quad \ldots \quad (26)$$

where $S_n = \sum_{k=1}^{K} V_k \exp(nj\,\psi_k)$, and is the signal output of the element k before the attenuator network. Using vector notation

$$y = \tilde{\mathbf{W}}\mathbf{S} = \tilde{\mathbf{S}}\mathbf{W} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (27)$$

the total power is given by $y^*\,y$ where

$$y^*\,y = \tilde{\mathbf{W}}^*\mathbf{S}^*\tilde{\mathbf{S}}\mathbf{W} = \tilde{\mathbf{W}}^*\mathbf{G}\mathbf{W} \quad \ldots \quad \ldots \quad (28)$$

The total power output of the adaptive array is therefore given by $\mathbf{W}^*\,\mathbf{G}\mathbf{W}$ where $\mathbf{G}$ is a square matrix formed by the product $\mathbf{S}^*\,\tilde{\mathbf{S}}$. The average value of this matrix is known as the covariance matrix. The diagonal terms represent the power at the individual elements, and the off-diagonal terms contain the cross correlations between the signals at the different element pairs.

The simplest way of minimising this total power expression is to set $\mathbf{W} = 0$. This is clearly a trivial solution in that not only is the interference minimised, but also the wanted signal.

To prevent the gain being reduced to zero, some form of constraint needs to be introduced. The simplest form of this is the constant-gain constraint introduced in eqn. 13 above. In this case, we have

$$\begin{cases} P_{TOT} = \tilde{\mathbf{W}}^*\mathbf{G}\mathbf{W} & \ldots \quad \ldots \quad \ldots \quad \ldots \quad (29) \\ \tilde{\mathbf{C}}\mathbf{W} = 1 = \tilde{\mathbf{C}}^*\mathbf{W}^* & \ldots \quad \ldots \quad \ldots \quad \ldots \quad (30) \end{cases}$$

To minimise the power subject to the constraint, we use the technique of Lagrange multipliers and minimise

$$\tilde{\mathbf{W}}^*\mathbf{G}\mathbf{W} + \lambda(1 - \tilde{\mathbf{C}}^*\mathbf{W}^*) \quad \ldots \quad \ldots \quad \ldots \quad (31)$$

which gives the result

$$\mathbf{W} = \lambda\mathbf{G}^{-1}\mathbf{C}^* \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (32)$$

where λ is a constant given by

$$\lambda = 1/(\tilde{\mathbf{C}}\mathbf{G}^{-1}\mathbf{C}^*) \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (33)$$

In this case, it is possible to evaluate the weight vector $\mathbf{W}$ directly in theory, given the covariance matrix $\mathbf{G}$ which is a property only of the signals, noise and aerials, and not of the weighting networks. Again this is not a full closed-loop system and is therefore not satisfactory for good null performance. Algorithms for minimisation of interference are considered in the following Section.

Algorithms

Various algorithms have been discussed at length in the literature.[1,2] Many of the methods discussed are applicable to both the total-power and interference-only minimisation. Techniques such as those which measure the covariance matrix are particularly applicable to the total-power situation. In this paper, detailed discussion is limited to those suited particularly to interference-only minimisation and it is assumed in each case that a constant-wanted-signal constraint has been applied.

Methods considered are:

(i) element at a time (Southwell[3]);
(ii) steepest gradient descent (Widrow *et al.*[2]);
(iii) matrix search methods (Butler matrix, Fletcher and Powell[4]);
(iv) null steering methods.

Element at a Time

This method is the simplest available. With the constant-gain constraint defined above, it is applicable to both total-power minimisation and interference-only minimisation. In this algorithm, each control A_i is stepped in sequence, minimising the error signal as a function of each control before moving on to the next. This is a particularly easy algorithm to implement.

Steepest-gradient Search

This method requires the measurement of the gradients with respect to all $2N$ variables. If it were possible to measure all these simultaneously, then this method would have a faster approach than above. It has also the advantage of tending to maintain the best signal-to-noise ratio.

Matrix Methods
Matrix methods are those in which there is a matrix relationship between the controls **A** (the independent variables) and the weights **W** (the dependent variables). The constant-gain facility is a particularly simple case of this where $\mathbf{\delta W} = \mathbf{M\delta A}$ and **M** is a fixed and easily defined matrix.

For a general matrix relationship, $\mathbf{W} = \mathbf{MA}$ where the A_i are the complex controls, and $\mathbf{T} = \mathbf{UW}_1$ so that

$$\mathbf{T} = \mathbf{UMA} \qquad (34)$$

Various matrices were considered and particular examples are:

(*a*) *Butler Matrix*. For this matrix, each control A_i adjusts the weights so as to adjust the aerial pattern mainly in a particular direction, and the matrix defines N search directions. The matrix is fixed and does not require prior knowledge of the direction of the interference.
(*b*) *Fletcher and Powell*. The matrix involved here relates the changes in weights to the steepest gradient vector **g** by the relationship

$$\mathbf{\delta W} = -\alpha\mathbf{Hg} \quad \text{where } \alpha = \text{constant} \qquad (35)$$

The constant α and the matrix **H** are calculated and updated within the algorithm. It is interesting to note that in the steepest gradient search, $\mathbf{\delta W} = -\alpha\mathbf{g}$ so that Fletcher and Powell differs in making its steps in directions not along the local line of steepest descent.
(*c*) *A Priori Search Algorithm*. The matrix here is chosen such that each control A_i adjusts the interference from one source only without affecting any other source. We have

$$\mathbf{\delta T} = \mathbf{U\delta W} = \mathbf{UM\delta A} \qquad (36)$$

From earlier, we note that it is necessary to pad the U matrix out to $N \times N$ with fictitious sources to be able to calculate a matrix M such that the matrix **UM** is diagonal. Then we obtain

$$\delta T_i = \lambda_i \delta A_i \qquad (37)$$

Null-steering methods
From eqn. 5, we have

$$T_k = V_k \sum_{n=1}^{N} W_n \exp(nj\psi_k) \qquad (38)$$

If we write

$$\exp(j\psi_k) = z_k \qquad (39)$$

then

$$T_k = V_k z_k \sum_{n=1}^{N} W_n z_k^{n-1} \qquad (40)$$

This can be written as a product of factors:

$$T_k = V_k z_k W_n \prod_{n=1}^{N-1} (z_k - Z_n) \qquad (41)$$

which has $(N-1)$ roots Z_n.

These roots can be represented as points in the complex plane as shown in Fig. 2. The real array pattern is represented on the unit circle ($|z| = 1$), so that roots of the polynomial which fall on the unit circle represent real nulls of the array pattern. The interferers or jammers can be represented by points z_k on the unit circle. A null is obtained for a particular bearing θ_k when there is at least one root Z_n equal to z_k. Furthermore, if for every value of z_k corresponding to an interfering source there is at least one root Z_n equal to z_k, then the interference is totally nulled.

A search can therefore be set up using the Z_ns as variables, seeking a minimum value of E. Since each T_k is itself a product function of the variables Z_n, E is no longer quadratic as for the weights or controls above, but is a polynomial and is therefore a considerably more complicated error function. It is interesting to note that, necessarily,

$$W_{n+1} W_N^* = W_1 W_{N-n}^* \quad \text{for all } n \qquad (42)$$

if the roots Z_n are to lie on the unit circle and correspond to real values of θ.

Although, for null steering, it is possible to use conventional weighting networks, the value for each weight is a polynomial function of all the roots. An alternative implementation is that of the Davies tree,[5]

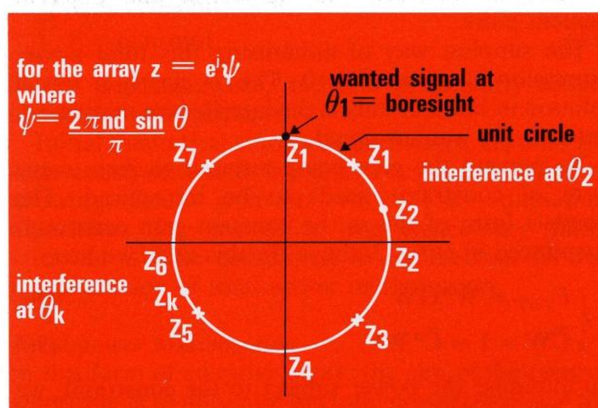

Fig. 2. Complex (Z) plane diagram in which the roots can be represented as points in the complex plane. Seven roots, Z_1 to Z_7 of an eight-element uniformly-weighted array are depicted.

in which each of the $N-1$ roots can be independently controlled directly. It is possible to incorporate constant-gain working into the null-steering method, but the theory is too lengthy to include here.

Signal/noise Ratio

Signal/noise ratio for the type of adaptive aerial discussed here can be considered as follows:

For an on-boresight wanted signal with an on-boresight constraint

$$V_{out} = V_{in} \sum_{n=1}^{N} (X_n + jY_n) = \text{constant} \quad \quad (43)$$

The noise contribution from each preamplifier is uncorrelated and equal so that the total noise power P is given by

$$P = p \sum_{n=1}^{N} (X_n^2 + Y_n^2)$$

$$= p \sum_{n=1}^{N} \{(X_n - \bar{X})^2 + (Y_n - \bar{Y})^2\} + \text{constant} \quad (44)$$

The noise therefore increases and the signal/noise ratio decreases as the controls X_n and Y_n deviate from the mean values which are used as the starting conditions for constant gain. A similar result also applies for the off-boresight constant-gain conditions given above.

Theoretical Conclusions

Analysis for a uniformly spaced array of N elements, for a wanted signal on boresight, shows that the ideal element spacing is $\sim\frac{2}{3}\lambda$, which combines reasonable directivity of the array with ease of control of the nulls. If the element spacing exceeds one wavelength, then there are non-zero real values of θ which correspond to $\psi = \pm 2\pi$, which therefore means that there are real directions in which interference cannot be nulled without nulling the wanted signal. If the wanted signal is off boresight, then the element spacing must be reduced to $\lesssim \lambda/2$.

There are several advantages in being able to recognise signal from interference:

(*a*) It is possible to recognise and therefore minimise the interference without affecting the wanted signal.

(*b*) Only a very simple constraint is required to maintain the wanted signal level.

(*c*) With cartesian control networks (i.e. $X_i + jY_i$) for the aerial elements, the interference is a simple quadratic function of the X_i and Y_i when a constant-wanted-signal constraint is applied. Under these conditions there is a wide range of optimisation algorithms which can be used, as the error function is well behaved and does not have any spurious minima.

(*d*) The system should be more rugged in its response to changes in wanted signal.

For television, for example, the wanted signal has synchronising pulses which can be clearly identified, so that any variations in pulse height must result from interference. For other signals this may be rather more difficult, although the advantages that result outweigh the pattern-recognition problems.

For computer simulations in the absence of noise, algorithms such as Fletcher and Powell or the preset search directions, give a far faster approach to the required minimum than for example the simple hill climb. Practical systems have three major sources of signal contributing to the error signal E: front-end and receiver noise, the independent interfering signals and cross correlation between interfering signals which appears in the form of beats between various carriers.

For the algorithms considered above, in the presence of noise, there is little advantage in terms of speed in using complex algorithms, particularly when operating close to the required minimum. The steepest gradient search tends to favour the best signal/noise ratio, and, if it were possible to measure all gradients simultaneously, would have an advantage in terms of speed. It also has been shown by computer simulation to have a more controlled behaviour when interfering signals are added or removed.

The null-steering algorithm has the particular disadvantage that the error surface can have mini-max points—i.e. points at which the gradient in all directions is zero, although the point is not a true minimum. This means that, if the starting conditions are not ideal, the algorithm cannot optimise correctly.

The limitations on adaptation rates caused by correlations and coherence between wanted signal and interference or jammers, and between interference signals themselves, have been studied both theoretically and practically and are considerable. The estimates of E for interference-only measurements, or of the total power or covariance matrix for total power measurements, can be significantly in error if the measurement time is shorter than the coherence time of the signals. Under these conditions, the array can fail to converge correctly to the required minimum. The theory and design of the measuring system is therefore an important part of adaptive-array design.

General Conclusions

This paper has presented a simple view of the theory of adaptive arrays and, in particular, of the linear uniformly spaced array of aerial elements. The work described here was directed particularly at the practical system described in the earlier sections of this volume, and therefore has, of necessity, not covered all aspects of adaptive arrays. For a more detailed coverage of specific aspects, the attention of the reader is drawn to review papers such as refs. 1 and 2 and also refs. 6, 7 and 8.

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Développements en Matière de Radio-Fréquences—Etude d'Introduction

Sommaire

Un laboratoire-RF situé au sein même du milieu de la radiodiffusion, couvre les applications des techniques relatives aux nouveaux circuits et au traitement des signaux avec pour objet de surmonter les limitations d'image et de qualité du son des émissions de télévision et de radio. La technique moderne a augmenté d'une façon significative le degré de sensibilité des récepteurs, mais, paradoxalement, a eu pour résultat d'accroître l'importance des effets des bruits et vacillement d'oscillateur et en particulier ceux des sources de fréquences synthétisées. Les distortions inhérentes des émissions sur bande latérale unique, lorsque celles-ci sont démodulées par un simple détecteur de groupe, ont engendré l'usage d'une démodulation synchrone dans les récepteurs et les équipements d'essai. Les ingénieurs-techniciens RF ont travaillé sans relâche à rechercher des méthodes pour minimiser les effets de diaphonie (CCI) engendrée par la propagation trans-horizontale anormale des signaux-UHF et à cet égard la présente section décrit un certain nombre des mécanismes résultant de cette propagation sur des voies maritimes et terrestres.

L'utilisation d'éléments de circuit et d'antenne pour combattre la diaphonie (CCI) et les propagations à voies multiples est également examinée. Un compte-rendu sommaire est élaboré concernant la mise au point du système à entourage sonore du type MSC, ainsi que le besoin pour un système de démultiplexeur amélioré à l'usage d'émissions à deux ou trois voies.

Antennes Adaptables pour Liaisons de Recepteurs de Retransmission—UHF

Sommaire

Le grand nombre sans cesse croissant d'émetteurs de télévision-UHF qui fonctionnent dans les bandes IV et V, a entraîné une augmentation du risque de diaphonie dans les liaisons de récepteurs de retransmission utilisés pour fournir des alimentations de programmes pour la plupart des émetteurs. A cet égard, des problèmes particulièrement difficiles se sont présentés aux planificateurs travaillant sur une liaison de récepteurs de retransmission de haute qualité sur la voie maritime de 135 km entre Stockland Hill, Le Devon et Alderney (Aurigny) l'île Anglo-Normande la plus proche de l'Angleterre. Afin de produire une solution du problème, les techniciens de la IBA ont développé une antenne directionnelle adaptable qui ajuste automatiquement et constamment son diagramme polaire et dans laquelle la restriction consiste à minimiser le niveau de la diaphonie en fonction du signal désiré. Cette étude examine les besoins fondamentaux et la conception de base des antennes directionnelles adaptables de nature complex ainsi que la mise en oeuvre d'un système construit sous forme de 2×2 rideaux de 8×2 antennes dipôles. Le rideau d'antenne fournit un gain de l'ordre de 24 dB et produit une performance de réjection à réglage continu d'environ 45 dB, soit quelques 20 dB de mieux que le résultat normalement obtenu même en utilisant un système d'antenne fixe de nature complexe. Le rideau d'antenne directionnelle doit rejeter les signaux indésirables à 7° seulement du relèvement du signal désiré. Certaines suggestions sont également formulées pour des antennes directionnelles adaptables plus simples à quatre éléments qui seraient tout à fait adéquates pour les emplacements difficiles, tels que par exemple les installations insulaires où les sources potentielles d'interférences parasitaires sont beaucoup nombreuses.

Techniques de Suppression Diaphoniques

Sommaire

Cette étude décrit un certain nombre de résultats en provenance d'une enquête détaillée des techniques de traitement de signaux permettant de faire une distinction contre la diaphonie 'décalée' à l'aide d'un système de filtrages à peigne adaptable. Le principe d'un filtrage à peigne d'un signal de télévision est basé sur le fait que l'énergie lumineuse est concentrée autour des harmoniques de la fréquence due balayage linéaire (15,625 kHz pour le Système-I). L'opération déréglée des émetteurs entraîne l'imbrication des deux spectres, mais ceux-ci peuvent néanmoins être séparés à l'aide d'un filtre à peigne, c'est à dire un filtre transversal avec lignes à retard de durée de ligne (64 μsec). Les principales caractéristiques d'un filtre du type expérimental sont décrites et la performance du dispositif est illustrée sous différentes conditions et modes opérationnels. Le filtre à peigne s'interface avec un récepteur en utilisant une démodulation synchrone et de plus permet d'effectuer l'annulation simultanée des porteuses parasitaires sur les deux décalages, conjointement avec les composants de luminance de basse-fréquence des signaux perturbateurs. L'étude montre que le système du filtrage à peigne est tout à fait praticable du point de vue technique et capable de produire une amélioration allant jusqu'à deux 'qualités d'image' (amélioration de l'ordre de 12 dB du taux de protection antidiaphonique) à condition toutefois que les porteuses indésirables soient 'décalées' par rapport à la porteuse désirée. Un filtre à peigne adaptable du type complexe qui utilise des signaux en-phase et déphasés peut réduire la diaphonie sur les deux 'décalages' tout en produisant une protection optimale à partir d'une seule source d'interférence. Le filtre peut être interfacé avec les récepteurs à démodulation synchrone et ne nécessite aucune modification de l'installation de l'antenne du récepteur de retransmission. Ainsi donc, le filtre à peigne fournit une autre arme anti-diaphonique dans l'arsenal de l'ingénieur chargé des conceptions de systèmes que vient complémenter les systèmes d'antennes adaptables plus complexes susceptibles d'être nécessaires lorsque des effets diaphoniques sévères sont probables ou que l'interférence parasitaire provient d'émetteurs 'non-décalés' par rapport à la fréquence porteuse.

Filtre à Peigne pour la Suppression Diaphonique des Signaux de TV

Sommaire

Cette étude décrit la performance pratique d'un prototype de filtre à peigne conçu et fabriqué conformément aux indications suggérées à la section précédente. Le filtre expérimental fut soumis d'une part à des essais de laboratoire en utilisant des signaux diaphoniques simulés engendrés par deux émetteurs-IF d'essai modifiés et d'autre part à un petit essai pratique sur place dans une base d'entretien mobile de la IBA où des observations subjectives furent relevées sous des conditions de diaphonie locale.

Ces essais devaient subséquemment confirmer la valeur de ce système de filtrage comme étant une méthode effective pour l'amélioration du taux de protection diaphonique tout en imposant néanmoins une pénalisation relative à une certaine perte de résolution verticale.

Ces essais tendaient également à suggérer que pour les usages sur place il serait désirable de simplifier et réduire le nombre des dispositifs de contrôle et à cet effet un filtre simplifié a été mis au point. L'auteur estime également qu'il serait peut-être possible d'améliorer le fonctionnement du filtre en utilisant des éléments de retard alternatifs, tels que par exemple un registre à décalage, une mémoire à accès direct ou un dispositif à couplage de charge.

Emetteurs, Récepteurs et Démodulateurs-UHF de Haute Performance pour les Mesures-RF de Télévision

Sommaire

Le grand nombre d'émetteurs-UHF non-surveillés construits et exploités par la IBA a engendré un besoin pour l'apport d'une nouvelle gamme d'émetteurs et d'équipements de démodulation d'essai de haute qualité grâce auxquels il serait possible de relever des mesures précises se rapportant à l'évaluation, la mise en service et l'entretien des émetteurs et récepteurs de retransmission. Afin de répondre à ce besoin la IBA a mis au point trois ensembles réglables, à savoir: un récepteur-UHF; un démodulateur et un émetteur d'essai de basse puissance. Les spécifications, conceptions et applications de ces trois équipements d'essai très avancés du point de vue technique sont décrites dans cette étude. Similairement, un nouveau type d'équipement couramment à l'étude et destiné aux équipes d'entretien mobiles est également présenté. Les caractéristiques particulières de cet équipement très complet sont essentiellement son poids extra-léger et l'usage d'un système de contrôle par micro-processeur. L'étude fait également ressortir la signification pratique du bruit de phase associé aux synthétiseurs de fréquence et en particulier les modulations de phases imprévues. Cet aspect fait l'objet d'une discussion plus détaillée à la Section suivante.

Bruit de Phase de Synthétiseur et son Effet sur les Systèmes de Radiodiffusion

Sommaire

L'usage sans cesse croissant des synthétiseurs de fréquence dans les équipements d'émission-TV, les récepteurs de retransmission et même certains récepteurs domestiques, présente de nombreux avantages. Malheureusement, la pureté spectrale de la sortie en provenance d'un synthétiseur de circuit bouclé à blocage de phase est nettement inférieure à celle d'un oscillateur à cristal. En effet, le bruit de phase beaucoup plus élevé a des effets pratiques qu'il convient de ne pas ignorer; le bruit de phase d'un oscillateur local est transféré à l'étage mélangeur, même sous forme de double équilibrage, directement vers le signal et en conséquence à pour effet de réduire la performance du système signal-bruit.

Cette étude présente une revue sommaire de la théorie du circuit bouclé à blocage de phase, tel qu'il est utilisé dans les synthétiseurs de fréquence-UHF et les démodulateurs synchrones, et montre les effets du bruit de phase à différents points à l'intérieur du circuit. Ceci mène à considérer la conception pratique des circuits bouclés à bruit faible ainsi que les mécanismes du processus de conversion par lesquels le bruit de phase apparaît sous forme de bruit sur les signaux-AF et vidéo. Les effets du bruit de synthétiseur dans réseau de radiodiffusion sont également examinés. Ainsi par exemple, le fait d'obtenir une performance signal-bruit acceptable à une émission radiodiffusée, fait appel à une spécification 'cible' de bruit de phase située dans la gamme des 40 Hz à 5,5 MHz avec 1° de crête-à-crête. L'étude démontre également que dans le cas des récepteurs du type domestique, basés sur des composants peu coûteux, les effets du bruit de phase seront peut-être considérablement plus sévères que pour les émissions radiodiffusées et à cet égard il est important que les concepteurs de récepteurs soient effectivement conscients du problème de l'interaction entre le bruit de phase, les modulations de phase imprévues et les démodulateurs synchrones.

Démultiplexeur Amélioré pour les radio-diffusions Stéréo où à 3-Voies

Sommaire

Faisant suite à plusieurs années d'études de recherche dans le domaine des systèmes et techniques de l'entourage sonore, les techniciens de la IBA ont mis au point un système matriciel à trois voies désigné par le symbole MSC (Demande de brevet déposée). Ce système, qui a été utilisé dans des radiodiffusions supra-atmosphériques expérimentales, représente le premier système à entourage sonore qui subjectivement parlant est totalement compatible avec les récepteurs-stéréo existants, ainsi qu'on peut le constater d'après le description de la Revue Technique No. 14 de la IBA.

Une importante partie de ce travail fut dirigée vers la conception de circuits peu coûteux suffisamment adéquats pour être utilisés dans des récepteurs domestiques de haute qualité. Cette mesure a eu pour effet de créer un nouveau démultiplexeur amélioré qui est tout à fait adéquat non seulement pour les radiodiffusions à entourage sonore de 3 voies, mais également sous forme de décodeur-stéréo de première qualité.

L'étude identifie certaines imperfections communes des décodeurs-stéréo du type conventionnel, y compris la production de sifflements parasitaires (bruits de gazouillement) provenant des émissions de voies adjacentes et de la perte d'énergie du rapport signal-bruit par suite du bruit de phase des circuits bouclés à blocage de phase.

Le démultiplexeur amélioré fait appel à des nouveaux démodulateurs qui éliminent la nécessité du filtrage 'anti-sifflement' ainsi qu'à un système à circuit bouclé de blocage de phase à deux modes qui permet de répondre effectivement aux exigences les plus sévères de la réception à 3 voies, sans pour cela nécessiter des ajustements préréglés de nature critique. Un principe de conception fondamental conjointement avec des détails de circuits de base sont présentés à l'usage d'une conception relativement simple qui est capable d'une part de produire une réception domestique convenable de radiodiffusions à émissions de 3 voies sans entraîner des ajustements de réglages critiques et d'autre part une excellente réception stéréophonique.

Appendice: Antennes Directionnelles Adaptables—Introduction Théorique

Sommaire

Une des section antérieures de cette *Revue Technique de la IBA* décrit la conception et la mise en opération pratiques des rideaux d'antennes directionnelles adaptables qui sont capables de s'ajuster automatiquement pour répondre à certaines contraintes spécifiques. Ces contraintes comprennent entre autres la réjection optimale des signaux indésirables qui causent des perturbations diaphoniques. Cet Appendice présente une introduction théorique rigoureuse relative aux systèmes d'antennes adaptables lorsque celles-ci sont basées sur des rideaux à éléments identiques, linéaires et à espacement uniforme, conjointement avec des algorithmes d'adaptation possibles. L'étude fait également ressortir le fait qu'il existe certaines limitations significatives sur les régimes d'adaptation engendrées par des corrélations et des cohésions entre les signaux désirables et indésirables (soit sous forme de diaphonie ou de brouillage électronique intentionnel). C'est pour cette raison que le principe et la conception du système de mesure représentent un élément important de la conception des rideaux d'antennes directionnelles adaptables pour une grande variété d'applications différentes.

Hochfrequenz-Entwicklungen — Eine einleitende Übersicht

Zusammenfassung

Im Rahmen des Rundfunkwesens befasst sich ein Hochfrequenzlabor mit der Anwendung neuer Schaltungs- und Signalverarbeitungsverfahren zur Überwindung von Beschränkungen der Bild- und Tonquatität bei Rundfunk- und Fernsehübertragungen. Die Empfindlichkeit moderner Empfangsgeräte hat sich durch die heutige Technologie wesentlich erhöht, jedoch hat dies dazu geführt, dass sich das Rauschen und Zittern von Oszillatoren, besonders bei Quellen mit Frequenzgeneratoren, wesentlich stärker bemerkbar machen als früher. Die Eigenverzerrung bei Restseitenbandübertragungen, die mit einem einfachen Hüllkurvengleichrichter demoduliert werden, haben zur Anwendung von Synchrondemodulation sowohl in Empfängern als auch Prüfgeräten geführt. Hochfrequenztechniker haben des weiteren grössere Anstrengungen gemacht, um die sog. Gleichkanalstörungen (CCI) zu minimieren, die durch transhorizontale abnormale Ausbreitung von UHF-Signalen verursacht wird. In diesem Abschnitt werden einige der Vorgänge beschrieben, die eine solche Ausbreitung auf Land- und Seeübertragungswegen verursachen.

Der Einsatz von Schaltungs- und Antennenelementen zur Überwindung von Gleichkanalstörungen und Mehrwegausbreitung wird betrachtet. Ausserdem wird eine kurze Übersicht über die Entwicklung des Dreikanal-Matrixsystems (MSC) und die Anforderungen für einen verbesserten Demultiplexor für Zwei- oder Dreikanalübertragungen vermittelt.

Selbstanpassende Antennen für UHF-Wiederausstrahlempfänger

Zusammenfassung

Die grosse und ständig weiter ansteigende Zahl von UHF-Fernsehsendern, die in den Bändern IV und V arbeiten, hat ein erhöhtes Risiko von Gleichkanalstörungen (CCI) der Wiederaustrahlempfänger (RBR) mit sich gebracht, die für die Programmversorgung der bei weitaus grössten Zahl von Sendern eingesetzt werden. Vor besonders schwierige Probleme sahen sich die Planer der Wiederausstrahlempfängerverbindung zwischen Stockland Hill in Devonshire und Alderney, der nächstgelegenen der Kanalinseln, bei der eine Seestrecke von 135 km überwunden werden muss, gestellt. Als Lösung entwickelten die Techniker der IBA (Independent Broadcasting Authority) eine Antennenanordnung, bei der die Polaritätscharakteristik laufend automatisch angepasst wird, wobei die Einschränkung die Minimierung der Gleichkanalstörung im Verhältnis zu dem gewünschten Signal ist. In dieser Abhandlung werden die grundsätzlichen Anforderungen und die Konstruktion komplizierter, selbstanpassender Anordnungen sowie die Realisierung einer als 2×2 Gruppierung von 8×2 Dipolen ausgebildeten Anordnung. Diese Anordnung liefert einen Verstärkungsgrad von ca. 24 dB und besitzt eine stufenlos veränderbare Unterdrückungsleistung von ca. 45 dB, d.h. ca. 20 dB über dem Wert, der sich unter normalen Bedingungen mit einer komplizierten, feststehenden Anordnung erreichen lässt. Diese Anordnung muss unerwünschte Signale unterdrücken, die nur 7° von der Peilung des gewünschten Signals abweichen. Des weiteren werden einige Vorschläge für vereinfachte selbstanpassende Anordnungen mit vier Elementen gemacht, die sich für schwierige Bereiche, einschlieslich von Inseln, eignen, wo die Anzahl der potentiellen Störquellen niedriger liegt.

Verfahren für die Unterdrückung von Gleichkanalstörungen

Zusammenfassung

In dieser Abhandlung werden einige Ergebnisse einer detaillierten Untersuchung der Signalverarbeitungsverfahren für die Unterdrückung von sog. 'versetzten' Gleichkanalstörungen (CCI) durch selbstanpassende Kammfilter beschrieben. Die Basis für die Kammfiltration eines Fernsehsignals liegt in der Tatsache, dass sich die Helligkeitsenergie um Oberschwingungen der Zeilenabtassfrequenz (bei System I 15,625 kHz) herum konzentriert. Eine versetzte Funktion der Sender verursacht eine Schichtung der beiden Spektren, jedoch ist eine Trennung mit Hilfe eines Kammfilters, d.h. einem Querfilter mit zeilenperiodischen Verzugszeilen (64 μs) möglich. Die Hauptmerkmale eines versuchsweise eingesetzten Filters werden zusammen mit der Leistung unter unterschiedlichen Einsatz- und Funktionsbedingungen beschrieben. Der Kammfilter hat eine Schnittstelle mit einem Empfänger, bei dem synchrone Demodulierung zur Anwendung kommt und erlaubt die gleichzeitige Aufhebung der störenden Trägerwellen in beiden Versatzrichtungen zusammen mit den niedrigfrequenten Hell Helligkeitskomponenten der Störsignale. Es wird dargelegt, dass der Kammfilter technisch durchführbar ist und in der Lage ist, Verbesserungen von bis zu zwei Bildqualitäten (ca. 12 dB Verbessurung im Gleichkanal-Schutzverhältnis) zu bewirken, sofern die unerwünschten Trägerfrequenzen gegenüber der gewünschten Trägerfrequenz 'versetzt' sind. Eine selbstanpassende Mehrfachkamm filteranordnung unter Einsatz sowohl von phasengleichen und Quadratursignalen kann die Gleichkanalstörung in beiden Versatzrichtungen reduzieren und gewährleistet gleichzeitig optimalen Schutz gegen eine Störquelle. Dieser Filter kann zusammen mit vorhandenen Synchrondemodulationsempfängerneingesetzt werden und erfordert keinen Umbau der Antennenanlagen von Wiederausstrahlempfängern. Der Kammfilter is demgemäss eine weitere Waffe im Waffenlager des Systemtechnikers beim Kampf gegen die Gleichkanalstörung und ergänzt die aufwendigen selbstanpassenden Antennensysteme, die bei besonders starken Gleichkanalstörungen oder dann erforderlich sind, wenn die Störung durch Sender verursacht wird, die in ihrer Trägerfrequenz nicht 'versetzt' sind.

Ein Kammfilter für die Unterdrückung von Gleichkanalstörungen von Fernsehsignalen

Zusammenfassung

In dieser Abhandlung wird die praktische Leistung eines Kammfilter-Prototyps beschrieben, der gemäss den Angaben im vorherigen Abschnitt konstruiert und ausgeführt wurde. Das Versuchsfilter wurde im Verlauf von Laboruntersuchungen unter Anwendung von mit zwei modifizierten ZF-Prüfsendern erzeugten, simulierten Gleichkanalstörsignalen und im Rahmen eines kurzen Versuchs unter Einsatzbedingungen auf einem ortsveränderlichen Einsatzwagen der IBA untersucht, um subjektive Beobachtungen bei tatsächlichen Gleichkanalstörungen zu ermöglichen.

Die Versuche bestätigten den Wert dieses Filteranordnung als Hilfsmittel zur Verbesserung des Gleichkanalstörverhaltens, obwohl ein gewisser Verlust der vertikalen Auflösung in Kauf genommen werden musste.

Aus den Untersuchungen unter Einsatzbedingungen ergab sich weiter, dass die Vereinfachung des Filters und eine verminderte Anzahl von Einstellelementen wünschenswert war, und demgemäss wurde eine vereinfachte Ausführung entwickelt. Der Verfasser ist der Auffassung, dass die Realisierung des Filters auch mit anderen Verzögerungselementen möglich ist, z. B. mit einem Schieberegister, einem Speicher mit wahlfreiem Zugriff oder einer ladungsgekoppelten Vorrichtung.

Hochleistungs-UHF-Sender, Empfänger und Demodulatoren für Fernseh-HF-Messungen

Zusammenfassung

Die grosse Anzahl von unbesetzten UHF-Sendern, wie sie von der IBA gebaut und betrieben werden, führte zur Notwendigkeit der Schaffung einer neuen Generation von Prüfsendern von hoher Qualität und entsprechenden Demodulationsausrüstungen, mit denen die Präzisionsmessungen durchgeführt werden können, die mit der Auswertung, Inbetriebnahme und der Instandhaltung von Sendern und Wiederausstrahlempfängern assoziiert sind. Zur Erfüllung dieses Anforderung entwickelte die IBA drei abstimmbare Ausrüstungsgruppen: UHF-Empfänger, Demodulator und Prüfsender mit niedriger Leistung. Die Spezifikation, Konstruktion und der Einsatz dieser drei ausgereiften Prüfausrüstungen werden beschrieben. Des weiteren werden neue Ausrüstungen vorgestellt, die für Einsatzmannschaften für die Instandhaltung bestimmt sind. Ein besonderes Merkmal dieser umfassenden Ausrüstungsreihe ist das geringe Gewicht und der Einsatz einer Mikroprozessorsteuerung. In der Abhandlung wird auch auf die praktische Bedeutung des Phasenrauschens, das mit Frequenzgeneratoren assoziiert ist, und besonders die Zwischenphasenmodulation hingewiesen. Eine nähere Erörterung dieses Themenkreises erfolgt im nachfolgenden Abschnitt.

Phasenrauschen von Frequenzgeneratoren und die Beeinflussung von Sendesystemen

Zusammenfassung

Der ständig wachsende Einsatz von Frequenzgeneratoren in Fernsehsendeausrüstungen, Wiederausstrahlempfängern und sogar in bestimmten Empfangsgeräten für den Hausgebrauch liefert viele Vorzüge. Unglücklicherweise ist die Reinheit des Ausgangsspektrums eines phasenverriegelten Schleifengenerators wesentlich niedriger als die eines Kristalloszillators.

Das wesentlich verstärkte Phasenrauschen hat praktische Auswirkungen, die nicht unbeachtet gelassen werden dürfen. Das Phasenrauschen des örtlichen Oszillators wird in der Mischstufe, selbst wenn es sich bei dieser um eine doppelsymmetrische Ausführung handelt, direkt auf das Signal übertragen und führt zu einer Verschlechterung des Rauschabstands im System.

In dieser Abhandlung wird kurz auf die Theorie der phasenverriegelten Schleife eingegangen, wie sie sowohl in UHF-Frequenzgeneratoren als auch in Synchrondemodulatoren eingesetzt wird, und es werden die Auswirkungen des Phasenrauschens auf die praktische Auslegung rauscharmer Schleifen sowie die Mechanik des Umwandlungsprozesses erörtert, bei dem das Phasenrauschen als Störpegel im Audio- oder Videosignal erscheint. Die Wirkung des Generatorrauschens in einer Senderanlage wird ebenfalls erörtert. Um beispielsweise einen Rauschabstand zu erzielen, der für eine Sendeanlage akzeptierbar ist, muss eine Zielspezifikation für Phasenrauschen im Bereich von 40 Hz bis 5,5 MHz von 1° Spitze-Spitze angestrebt werden. Es wird nachgewiesen, dass bei Geräten für den Hausgebrauch, die auf billigen Bausteinen basieren, die Wirkung des Phasenrauschens wesentlich ausgeprägter sein kann als bei Sendeanlagen, und Konstrukteure von Empfängers sollten daher die Wechselwirkung zwischen Phasenrauschen, Zwischenphasenmodulation und Synchrondemodulatoren unbedingt berücksichtigen.

Ein verbesserter Demultiplexer für Stereosignale bei Dreikanal-Matrixsystemen

Zusammenfassung

Nach mehrjähriger Forschung auf dem Gebiet von Raumtonsystemen und -verfahren haben die Techniker der IBA ein verbessertes Dreikanal-Matrixsystem entwickelt, das unter der Bezeichnung MSC (Patent angemeldet) geführt wird. Dieses System, welches bereits in Versuchssendungen zur Anwendung kam, ist das erste Raumtonsystem, dass rein subjektiv mit vorhandenen Stereoempfängern gemäss IBA Technical Review Nr. 14 kompatibel ist.

Ein wichtiger Bestandteil der Arbeiten auf diesem Gebiet war die Suche nach der Entwicklung preisgünstiger Schaltungen für den Einsatz in Qualitäts-Haushaltsempfängern. Dies führte zur Entwicklung eines neuen und verbesserten Demultiplexers, der sich nicht nur für Dreikanal-Raumtonsendungen, sondern auch als überlegener Stereodekoder geeignet ist.

In dieser Abhandlung werden häufige Schwächen konventioneller Stereodekoder identifiziert, darunter das Vorhandensein von sog. 'Vogelgeräuschen' (d.h. Zwitschergeräuschen), die durch Nachbarkanalsendungen und Verschlechterung des Rauschabstandes durch das Phasenrauschen von phasenverriegelten Schleifen verursacht werden.

Der verbesserte Demultiplexer besitzt neuartige Demodulatoren, die die Notwendigkeit einer Filtration der Vogelgeräusche überflüssig machen, sowie ein doppeltwirkendes, phasenverriegeltes System, das die schärferen Anforderungen für einen Dreikanalempfang ohne die Notwendigkeit von kritischen Voreinstellgliedern erfüllt. Es werden die theoretischen Grundregeln mit Einzelheiten der wesentlichen Schaltkreise für eine relativ einfache Ausführung dargelegt, mit der guter Heimempfang von Dreikanalsendungen ohne kritische Voreinstellung möglich ist, wobei der Empfänger gleichzeitig hervorragenden Stereoempfang zulässt.

Anhang: Selbstanpassende Antennenanordnungen—eine theoretische Einführung

Zusammenfassung

In einem vorhergehenden Abschnitt dieses *IBA Technical Review* werden die Konstruktion und die praktische Realisierung von selbstanpassenden Antennenanordnungen beschrieben, die sich automatisch anpassen, um bestimmten Anforderungen gerecht zu werden. Zu diesen Anforderungen gehört u.a. die optimale Unterdrückung unerwünschter Signale, die eine Gleichkanalstörung (CCI) verursachen können. Dieser Anhang enthält eine kritische theoretische Betrachtung von selbstanpassenden Antennenanordnungen, die auf der Basis von linear in gleichmässigen Abständen angeordneten, identischen Elementen aufgebaut sind, sowie mögliche Algorithmen für die Anpassung. Es wird dargelegt, dass durch Korrelation und Kohärenz zwischen den gewünschten und den unerwünschten Signalen die Anpassungsgeschwindigkeit erheblich beschränkt werden kann (und zwar unabhängig davon, ob es sich um Gleichkanalstörungen oder willkürliche elektronische Störsendungen handelt). Aus diesem Grunde ist die Theorie und die Konstruktion des Messsystems ein wichtiger Bestandteil der Konstruktion von selbstanpassenden Anordnungen für eine Vielfalt von Einsatzmöglichkeiten.

Desarrollos en Radiofrecuencia—Examen Preliminar

Resumen

El laboratorio de radiofrecuencia, dentro del ámbito de la radiodifusión, se ocupa de la aplicación de nuevas técnicas de circuitos y de proceso de señales para superar las limitaciones de la calidad de la imagen y del sonido de las transmisiones de televisión y radio. La tecnología moderna ha aumentado considerablemente la sensibilidad de los receptores pero, paradójicamente, ha dado lugar a que los efectos del ruido y las fluctuaciones del oscilador, especialmente los de las fuentes de frecuencia sintetizada, sean más importantes. La inherente distorsión de las transmisiones de banda lateral residual al ser demoduladas por un sencillo detector de envolvente, ha conducido al empleo de la demodulación síncrona en receptores y equipos de prueba. Los ingenieros de RF han trabajado mucho para tratar de reducir a un mínimo la interferencia común (CCI) originada por la propagación anómala de transhorizonte de las señales de UHF, describiéndose en esta sección algunos de los mecanismos que resultan en tal propagación sobre trayectos de mar y tierra.

Se considera el uso de elementos de circuitos y antenas para combatir la CCI y la propagación de caminos múltiples. Se da una breve relación del desarrollo del sistema de sonido circundante MSC y de la necesidad de un demultiplexor mejorado para transmisiones de dos o tres canales.

Antenas Adaptables para Enlaces RBR de Frecuencias Ultraelevadas

Resumen

El gran y siempre major número de transmisiones de televisión de frecuencias ultraelevadas (UHF) que operan en las bandas IV y V ha traído consigo un creciente riesgo de interferencia común (CCI) en los enlaces de receptores de redifusión (RBR) que se emplean para proporcionar alimentación de programas a la inmensa mayoría de los transmisores. Los problemas que se plantearon en el diseño del enlace RBR de alta calidad sobre el trayecto marino de 135 km entre Stockland Hill, Devon y Alderney, la más cercana de las islas del Canal de la Mancha fueron muy considerables. Para solucionarlos los ingenieros de IBA desarrollaron una red directiva de antenas adaptable que ajusta automática y continuamente su diagrama polar, habiendo de reducirse a un mínimo el nivel de CCI relativo a la señal deseada. Este artículo discute los requisitos básicos y el diseño de conjuntos adaptables complejos y el establecimiento de un sistema construido como conjunto 2×2 de 8×2 dipolos. El conjunto proporciona una ganancia de unos 24 dB con un rechazo de ajuste continuo de unos 45 dB, que es del orden de 20 dB mejor que el que puede alcanzarse empleando un conjunto complejo fijo. El conjunto tiene que rechazar señales indeseables solamente a 7° de la marcación de la señal deseada. Se hacen algunas sugerencias para conjuntos adaptables más sencillos de cuatro elementos que serían adecuados para lugares difíciles, incluyendo emplazamientos insulares, en que las fuentes potenciales de interferencia son menores.

Técnicas de Supresión de CCI

Resumen

Este artículo presenta algunos resultados de una detallada investigación sobre técnicas de proceso de señales para discriminar contra interferencia común 'descentrada' (CCI) por medio de filtrado de peine adaptable. El fundamento para el filtrado de peine de una señal de televisión radica en el hecho de que la energía de luminancia se concentra alrededor de los armónicos de la frecuencia de la línea de exploración (15,625 kHz para el sistema I). La operación descentrada de los transmisores hace que se intercalen los dos espectros, pero pueden ser separados por medio de un filtro peine, o sea, un filtro transversal con líneas de retardo de período lineal (64 μseg). Se describen las características principales de un filtro experimental, indicándose sus propiedades bajo distintas condiciones y modos de operación. El filtro peine se combina con un receptor empleando demodulación síncrona, permitiendo la cancelación simultánea de las portadoras que interfieren en ambos descentramientos, junto con los componentes de luminancia de baja frecuencia de las señales de interferencia. Se demuestra que el filtrado de peine es técnicamente factible y capaz de proporcionar un mejoramiento de hasta dos 'grados de imagen' (unos 12 dB de mejora en relación de protección cocanal) siempre que las portadoras indeseadas estén descentradas con respecto a la portadora deseada. Un filtro peine adaptable complejo, empleando señales en fase y en cuadratura puede reducir la CCI en ambos 'descentramientos' y proprocionará también óptima protección de una fuente de interferencia. El filtro puede interconectarse con receptores demoduladores síncronos existentes y no implica cambio alguno de la instalación de antena RBR. Así pues, el filtro peine proporciona un arma anti-CCI adicional para el arsenal del ingeniero diseñador de sistemas, complementándose con los más complicados sistemas de antenas adaptables que puedan necesitarse cuando sea probable una seria CCI o cuando la interferencia sea de transmisores que no estén 'descentrados' en frecuencia portadora.

Filtro Peine para la Supresión de CCI en Señales de Televisión

Resumen

Este artículo describe el funcionamiento práctico de un prototipo de filtro peine diseñado y construido según las líneas indicadas en la sección precedente. El filtro experimental fue sometido tanto a pruebas de laboratorio, empleando señales de CCI simuladas generadas por dos transmisores de prueba de frecuencia intermedia (IF) modificada, como también a una breve prueba de campo en una base de mantenimiento móvil de IBA en que se realizaron observaciones subjetivas en condiciones de campo de CCI.

Estas pruebas confirmaron el valor de tal filtrado como medio de mejorar la relación de protección de CCI, aunque con la desventaja de cierta pérdida de resolución vertical.

Estos ensayos indicaron también que para empleo en el campo sería aconsejable simplificar y reducir el número de controles, habiéndose desarrollado un filtro simplificado. El autor opina también que sería posible completar el filtro empleando elementos de retardo alternativos, por ejemplo un registrador de desplazamiento, una memoria RAM o un dispositivo de carga acoplada.

Transmisores de Ultra Frecuencia de Alto Rendimiento, Receptores y Demoduladores para Medidas de RF en Televisión

Resumen

El gran número de transmisores de ultra frecuencia no atendidos construidos y operados por la IBA crearon la necesidad de una serie de transmisores de prueba y equipo de demodulación de alta calidad con el que llevar a cabo las medidas de precisión asociadas con la evaluación, puesta en marcha y mantenimiento de transmisores y receptores de RBR. Para cumplir este requerimiento la IBA desarrolló tres

equipos sintonizables: receptor de ultra frecuencia; demodulador; y transmisor de prueba de baja potencia. Se describe las características, diseño y aplicación de estos tres complejos equipos de prueba. Se presenta un nuevo equipo diseñado generalmente para su empleo por los grupos de mantenimiento móviles. Como características especiales de este instrumento comprensivo pueden mencionarse su ligero peso y el uso de control por microprocesador. El artículo destaca también el significado práctico del ruido de fase asociado con los sintetizadores de frecuencia, especialmente la modulación de fase incidental. Esto se discute con más detalle en la sección siguiente.

Ruido de fase del Sientetizador y su Efecto en los Sistemas de Radiodifusión

Resumen

El creciente empleo de sintetizadores de frecuencia en equipos transmisores de televisión, receptores RBR e incluso algunos receptores domésticos, tiene muchas ventajas. Desgraciadamente, la pureza espectral de la salida de un sintetizador de circuito de fijación de fase es considerablemente inferior a la de un oscilador de cristal. El ruido de fase, muy superior, tiene efectos prácticos que no deben ignorarse; el ruido de fase de un oscilador local se transfiere en una etapa mezcladora, incluso en forma doblemente equilibrada, directamente a la señal y, por consiguiente, degrada la relación de señal a ruido del sistema.

El presente artículo revisa brevemente la teoría del circuito de fijación de fase según se emplea tanto en los sintetizadores de frecuencia de UHF como en los demoduladores síncronos, mostrando los efectos del ruido de fase en diversos puntos del circuito. Esto conduce a la consideración del diseño práctico de circuitos de bajo ruido y los mecanismos del proceso de conversión por el que el ruido de fase aparece como ruido en las señales de audio y video. Se discuten los efectos del ruido del sintetizador en una red de difusión. Por ejemplo, para conseguir una relación de señal a ruido aceptable para la radiodifusión se necesita alcanzar un ruido de fase en la banda de 40 Hz a 5,5 MHz de 1 grado pico a pico. Se demuestra que para receptores domésticos basados en componentes de bajo costo, los efectos del ruido de fase pueden ser mucho más serios que para la radiodifusión, y los diseñadores de receptores deberán advertir la interacción entre el ruido de fase, la modulación de fase incidental y los demoduladores síncronos.

Demultiplexor Mejorado para Difusiones Estereofónicas o de Tres Canales

Resumen

Después de varios años de investigaciones de sistemas y técnicas de sonido circundante, los ingenieros de IBA desarrollaron un sistema mejorado de matriz de tres canales denominado MSC (patente pendiente). Este sistema, que ha sido utilizado en radiodifusión experimental, es el primer sistema de sonido circundante enteramente compatible subjetivamente con los receptores estereofónicos existentes, como se describe en la IBA Technical Review No. 14.

Una parte importante de este trabajo fue destinada a producir circuitería de bajo costo adecuada para emplear en receptores domésticos de alta calidad. Así se ha conseguido un nuevo y mejorado demultiplexor adecuado no sólo para difusión de sonido circundante de tres canales, sino también como un decodificador estereofónico de excelente calidad.

El artículo identifica las limitaciones comunes en los decodificadores estereofónicos corrientes, incluyendo la producción de 'birdies' espúreos (o sea, ruidos de gorjeo) procedentes de transmisiones de canal adyacente y la degradación de la relación de señal a ruido debida al ruido de fase de los circuitos de fijación de fase.

El demultiplexor mejorado comprende nuevos demoduladores que evitan la necesidad del filtrado 'anti-birdie' y un sistema de circuito de fijación de fase de doble modo que cumple con los requisitos más exigentes de la recepción de tres canales sin necesidad de ajustes críticos preestablecidos. Se presenta la teoría de diseño fundemental junto con los detalles de circuito básicos para un diseño relativamente sencillo que es capaz de buena recepción doméstica de difusión de tres canales sin ajuste de fijación crítico, obteniéndose asimismo una excelente recepción estereofónica.

Apéndice: Conjuntos de Antenas Adaptables—Introducción Teórica

Resumen

Una de las secciones anteriores de esta *Revista Técnica IBA* describe el diseño y establecimiento prácticos de redes directivas de antenas autoadaptables que son capaces de ajustarse a sí mismas automáticamente para cumplir exigencias determinadas. Tales requerimientos comprenden el rechazo óptimo de señales indeseadas que producen interferencias comunes (CCI).

Este apéndice presenta una rigurosa introducción teórica a las antenas adaptables basadas en un conjunto de elementos idénticos lineales dispuestos uniformemente, junto con posibles algoritmos de adaptación. Se demuestra que existen limitaciones notables en los grados de adaptación debidos a correlaciones y coherencia entre las señales deseadas e indeseadas (lo mismo en forma de CCI que de interferencia electrónica intencionada). Por este motivo, la teoría y el diseño del sistema de medida constituyen una parte importante del diseño del conjunto adaptable para una gran diversidad de aplicaciones.

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